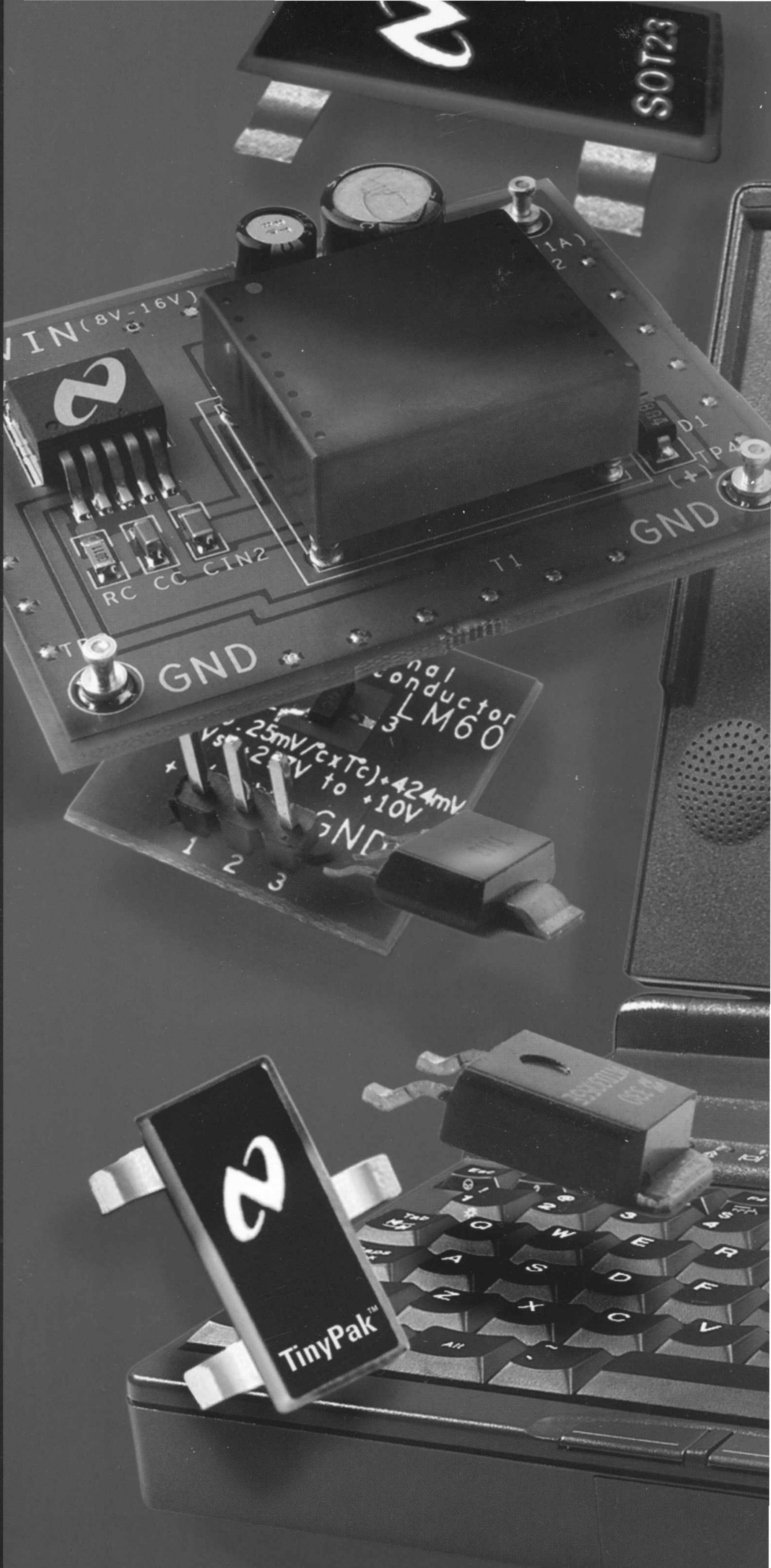
NATIONAL LINEAR SEMINAR 1996



National Semiconductor®

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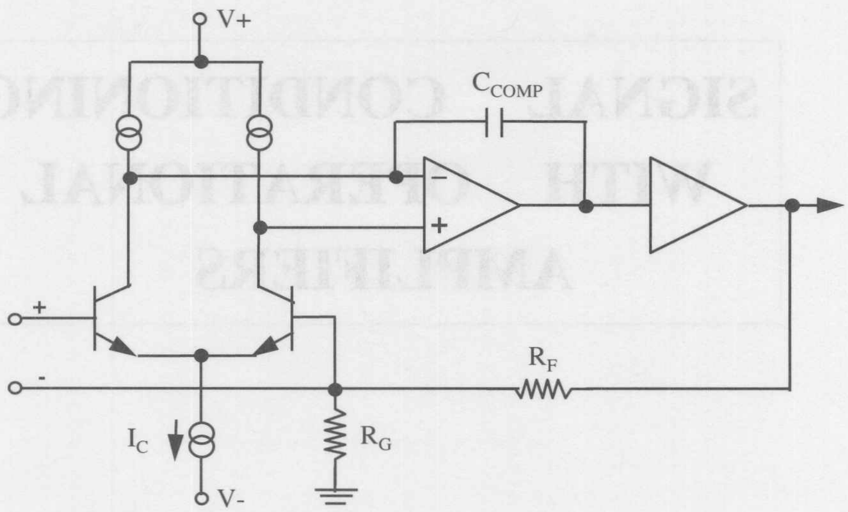
SIGNAL CONDITIONING WITH OPERATIONAL AMPLIFIERS

CONVENTIONAL OP AMP



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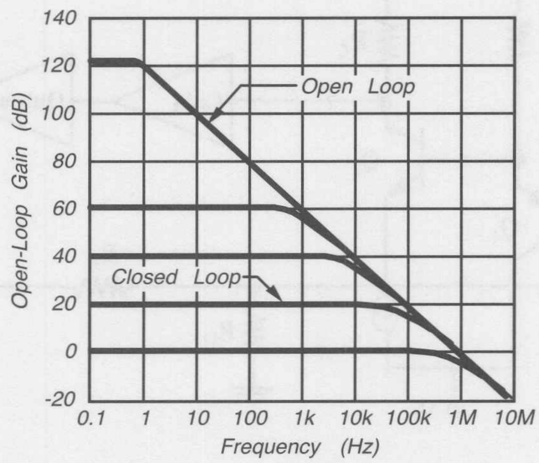
This simplified op amp schematic shows the major elements of a voltage feedback op amp.

The differential, high gain, common emitter input stage is followed by a differential to single ended gain stage and a unity gain output stage. Since the output stage is a unity gain follower, the amplifier open loop ac gain is set by the product of the input stage transconductance gm and the capacitor impedance. The closed loop gain is set by the ratio of the feedback resisters R_F and R_G. Ccomp is selected to reduce the open loop gain to unity before the phase shift through the amplifier has reached 180 degrees. If this were not the case, then a feedback network for low or unity gain (R_F approaching 0Ω) would cause the amplifier to oscillate. Since the impedance of C_{comp} decreases with frequency at a rate of 6dB/octave, the open loop gain of the amplifier will also decrease with frequency. C_{comp} also determines the maximum slew rate of the amplifier. The current available to charge or discharge the capacitor is the input stage current Ic, and this will determine the maximum rate of change, or slewing, of the output.

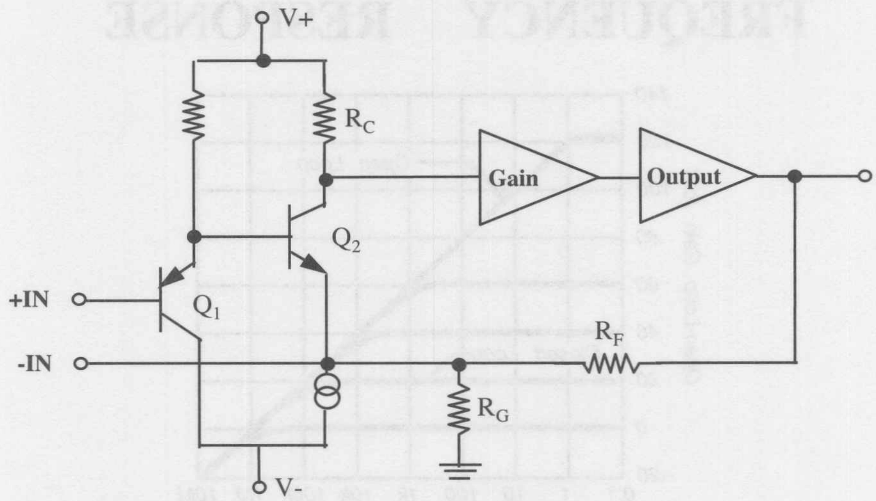
CONVENTIONAL VOLTAGE FEEDBACK OP AMP'S FREQUENCY RESPONSE



This single pole compensation produces the familiar response curves shown above. This response is typical of all unity gain stable voltage feedback op amps. Note that the limit of closed loop upper frequency response is the same as the open loop gain curve. This limits the bandwidth of this op amp to less than 100kHz when used at a closed loop gain of 10 or 20 dB. Increasing the closed loop gain reduces the bandwidth even further.

The position of the open loop gain slope is well defined by the input stage transconductance and the size of the capacitor. Both these parameters are well controlled in monolithic processes, ensuring stable unity gain operation of the amplifier. The frequency at which the gain starts to roll off depends also on the DC gain which is the product of the input stage transconductance and the second stage DC gain. Higher second stage DC gain will result in lower frequencies at which the ac gain starts to roll off rather than in a change in the unity gain crossover frequency.

CURRENT FEEDBACK AMPLIFIER



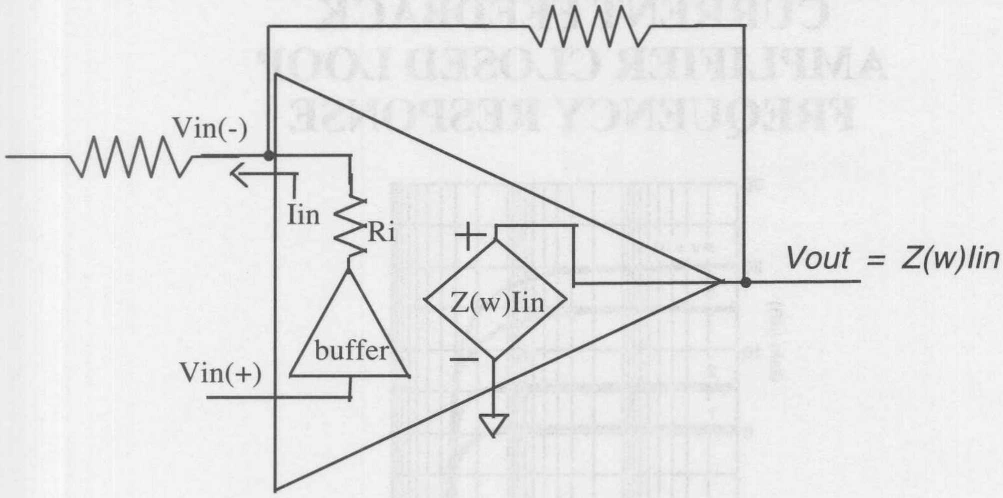
The current feedback amplifier has a completely different input stage from the voltage feedback op amp.

In this example, the non-inverting input, Q1, is connected as an emitter follower with no voltage gain. This connection eliminates Millet effect and has a very low output impedance. The signal is directly connected to the inverting input through the base-emitter of Q2. This direct, low impedance feedback to the inverting input results in much less phase shift than that which is present in the much longer path of a voltage feedback op amp. The open loop gain of the amplifier is provided by Q2 and a subsequent voltage gain stage. An output buffer provides a low impedance to drive the load and the feedback network that is setting the closed loop gain.

One result of the short feedback path is stability over a wider bandwidth when the loop is closed. This feature alone would result in an op amp with a wider bandwidth. However, the current feedback amplifier acts quite differently to a voltage feedback amplifier when the closed loop gain is adjusted by changing the ratio of the feedback resistors R_F and R_G.

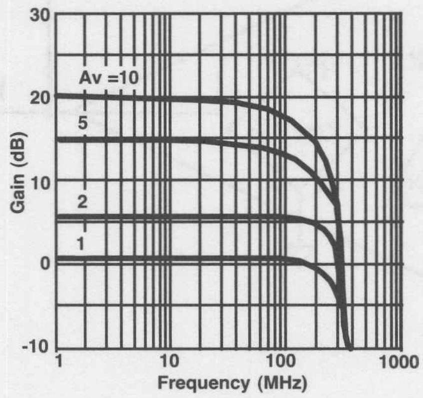
The current feedback key advantage comes about because the current level in the second stage is not only set by the current source in the emitter, but also by the shunting effect of the feedback resistor network. If R_F is kept constant, and R_G is reduced to raise the closed loop gain, the smaller value of R_G will raise the gain of Q2 correspondingly. The overall effect of raising the open loop gain as the closed loop gain is increased is to produce a nearly constant bandwidth.

Current Feedback Amplifier



The input of this amplifier takes Iin, the current difference between the non-inverting input Vin(+) and the inverting input Vin(-) and amplifies this signal by Z(w), the transimpedance gain of the current feedback amplifier. There is a buffer between the inverting and non-inverting inputs of current feedback amplifiers. The inverting input is low impedance because it is the output of the buffer. The non-inverting input is high impedance because it is the input of the buffer. Due to the different input impedances in CFA, the input bias currents of CFA are very different, and CFA datasheets specify the input bias currents of each node individually.

CURRENT FEEDBACK AMPLIFIER CLOSED LOOP FREQUENCY RESPONSE



The simultaneous increase in input stage gain, with increase in the closed loop gain, results in the set of closed loop gain curves shown above. Notice that the bandwidth has only decreased slightly while the closed loop gain is increased from 1 to 5. If the closed loop gain is increased even further, some reduction in bandwidth will become apparent, as can be seen in the gain of 10 curve.

Because the bandwidth does not change in direct proportion to the closed loop gain, GBW for a current feedback amplifier is not a standard specification. The effective GBW will increase as the closed loop is increased. This is because the amplifier open loop gain has increased, while the compensation has not.

LM7121 - Tiny, Low Power Voltage Feedback Amplifier

Features

- Voltage feedback topology
- Speed - Power tradeoff
  - Unity gain frequency of 105Mhz with Isupply of 5.1mA (Vs = ±5V)
- Stable with unlimited capacitive load
- Tiny SOT-23 package

The LM7121 is a high performance operational amplifier which addresses the AC performance needs of video and imaging applications, and the size and power constraints of portable applications. The LM7121 operates over a wide range of supply voltages, from a single 5V supply up to ±15V. Note that the part is specified for ±5V max in the SOT23-5 package. It offers excellent speed-power tradeoff delivering 1300V/ms and 235MHz bandwidth.

Two other features are the SOT23-5 package and amplifier stability while driving unlimited capacitive loads. The benefits of the package are evident in designs where space and weight are the critical parameters. The Tiny package allows the user to place the amplifier anywhere on a board close to the signal source or the next stage in the system; eg. an A/D converter.

Applications

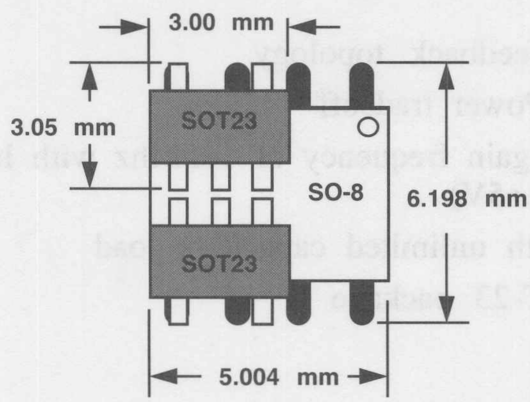
Cable Drivers

Set-top boxes

PC Video cards

Why use TinyPak™ devices ?

SOT23-5 TinyPak™ vs. SO8 to Scale



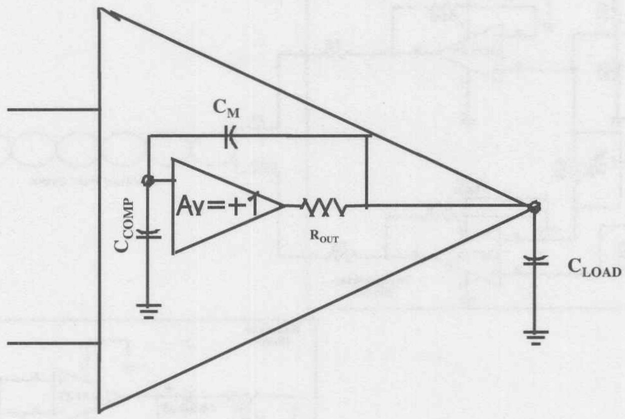
Why should you use TinyPak™ devices ?

There are many reasons.

They take less space on your board. A standard 8 pin molded DIP uses 10.34 times as much area as the SOT23 and 40 times as much volume. The SO-8 uses 3.6 times as much area and 5 times the volume. When boards are stacked on top of each other, the height is as important as the area.

The SOT-23 package eliminates the board layout hassle that results when you try to route the traces from all over the board to a dual or quad opamp. Those long traces make great antennas for picking up noise as well as using even more board space.

Unlimited Capacitive Load Driving Capability

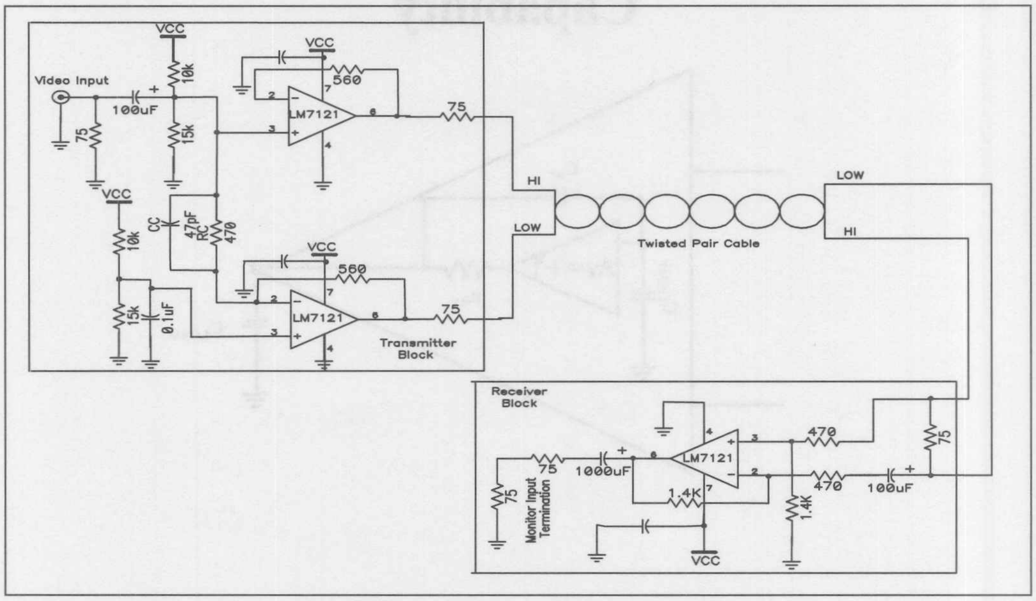


The LM7121, is capable of driving heavy capacitive loads without oscillating.

The capacitor, C_{COMP}, is the internal compensation capacitor tied to a high impedance node inside the opamp. The functional block, Av=+1, is the internal output stage with an output impedance Rout. As long as the voltage across C_{LOAD} and C_{COMP} track each other, the internal capacitor, C_M, is boot-strapped and has no effect on the operation of the amplifier. With large enough capacitive loads and at high enough frequencies, C_{LOAD} voltage will lag with respect to the C_{COMP} voltage and a portion of C_M will in effect appear to be in parallel to C_{COMP}.

In this fashion, the amplifier compensation becomes a dynamic parameter which varies with the amount of capacitive loading; as the capacitive load increases, the effective compensation increases with it in order to insure stable operation under all loads.

Video on Twisted Pair Cable



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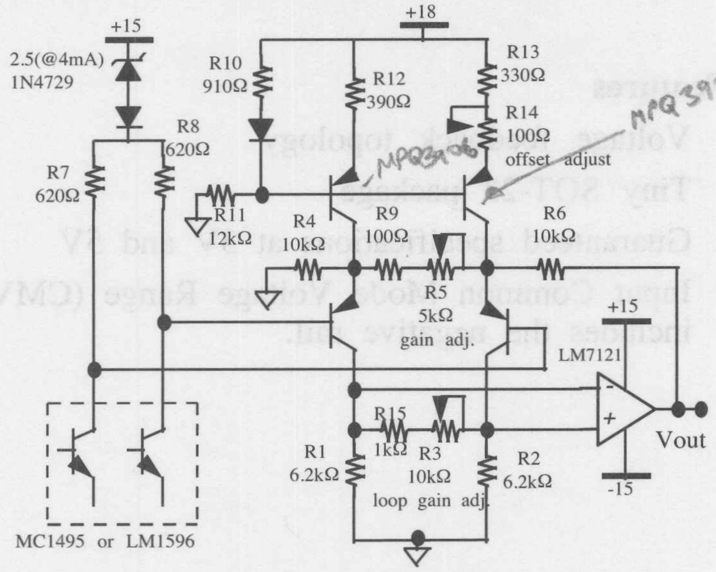
A very cost effective way of transmitting composite color video is to use twisted pair cables. This not only reduces wiring cost, it allows the designer to use the existing phone lines as the transmitting media but these cables exhibit a relatively high capacitance. The scheme shown above does that using a single power supply. The signal path is AC coupled in order to bias the LM7121 inputs close to the middle of its supply rails (Vcc and ground). Even though the characteristic impedance of a twisted pair is around 600 Ohms, the cable is terminated in 75 Ohms to take advantage of the lower impedance in order to reduce the RC time constant of the cable capacitance and termination resistance. The LM7121 is capable of driving composite video into a 75 Ohm back terminated cable. This makes the LM7121 a good choice for this application. For operation above Vcc of 10V, the DIP or SO package is recommended. When tested with 25 meters of cable, the output has less than 1% differential gain error and less than 1 degree differential phase error.

In the schematic shown, Rc and Cc can be adjusted to compensate for different wire lengths.

Differential to Single Ended Amplifier



The circuit shown above is ideal for converting a differential output signal with a high common mode voltage (such as the output of LM1596 modulator-demodulator or MC1495 multiplier) to a single-ended output referenced to ground. The advantages of this circuit over the traditional single opamp schemes are the following:

a) Gain is adjustable over a wide range with a single resistor.

b) Higher realizable gain bandwidth product (2.5Ghz with the gain set to 190).

Some key features are:

a) Unlimited capacitive load drive capability. This is one of the most distinguishing features of the LM7121 opamp.

b) Output slew rate of 700V/us.

c) Output Swing= 25V peak to peak.

d) Full power bandwidth= 8Mhz.

e) Variable gain from 5 to 201 (for values shown).

f) Bandwidth: 13Mhz with gain set to 190

35Mhz with gain set to 5

In the schematic above, R3 (Loop Gain) can be adjusted to set the damping factor of the step response .

LM7131 - Tiny High Speed Single Supply Amplifier

Features

- Voltage feedback topology
- Tiny SOT-23 package
- Guaranteed specifications at 3V and 5V
- Input Common Mode Voltage Range (CMVR) which includes the negative rail.

The LM7131 is ideal for portable applications due to its availability in the SOT23-5 package. Single supply 3V and 5V performance makes the amplifier ideal for driving A/D converters. The LM7131 can also be used in a wide variety of portable video and computing applications due to high CMRR (110dB typical) and PSRR (95dB typical). .

Applications

Drivers for A/D converters

4V output swing with +5V supply

High Fidelity digital audio

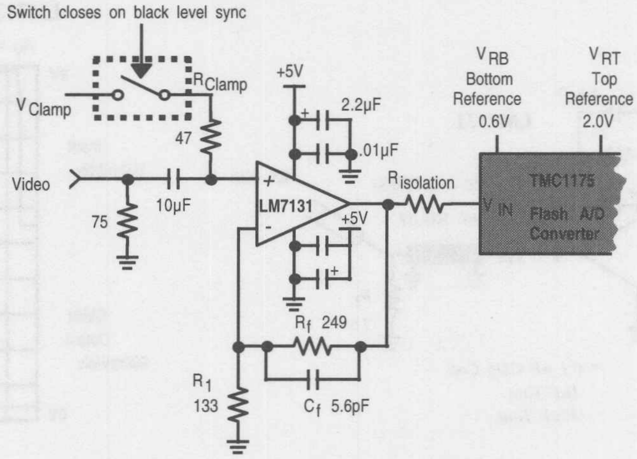
Video cards

Differential gain of 0.25%

Differential phase of 0.75degrees

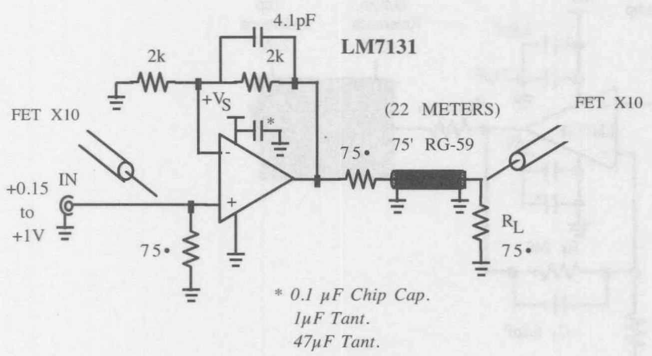
Total harmonic distortion of 0.1% at 4MHz

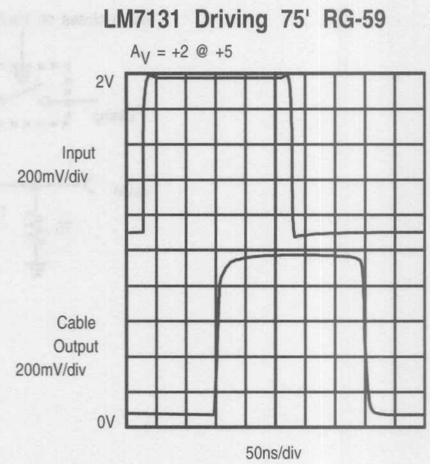
LM7131 Driving Flash A/D Converters



For a flash analog-to-digital converter, the converter input is usually a flying capacitor which switches between the input and an internal comparator ladder. When the capacitor switches to the input, it will have a voltage which is different than the op amp output. This will disturb the op amp output. To drive the A-to-D converter accurately, the op amp needs to provide high current to drive the capacitor toward the correct voltage, and settle to the correct final value.

LM7131 Driving 75' RG-59





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The LM7131 works well as a cable driver, especially in the SO-8 and DIP packages, which have better power dissipation than the SOT23-5 package. With a 5 volt single supply, the LM7131 can drive 4.0 volts across a 150 ohm load. It can also be used to drive back terminated 50 ohm cables. The delay in the bottom waveform is caused by the propagation time down the cable.

LM6171/LM6172 - Low Distortion Voltage Feedback Amplifiers

Features

- Single - LM6171, Dual - LM6172
- Voltage feedback topology
- Speed - Power tradeoff
  - Unity gain frequency of 100Mhz with Isupply of 2.5mA/channel
- Excellent differential gain and differential phase characteristics

The LM6171/LM6172 is built on National's advanced VIP™ (Vertically integrated PNP) complementary bipolar process. The LM6171/LM6172 is a high performance voltage feedback amplifier which offers a slewrate of 3600V/ms, and a unity-gain bandwidth of 100MHz while consuming only 2.5mA supply current.

Applications

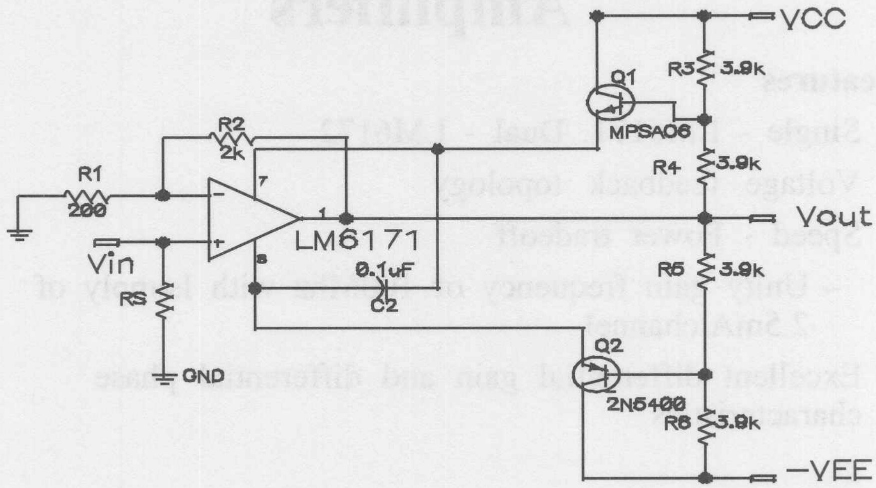
Video Amplifiers for NTSC, SECAM systems

HDTV Amplifiers

Pulse Amplifiers

The ±15V power supplies allow for large signal swings and give greater dynamic range in many applications. High output current, low SFDR (Spurious Free Dynamic Range) and THD (Total Harmonic Distortion) make the LM6171/LM6172 ideal for ADC/DAC systems. The impressive AC and DC performance of the LM6171/LM6172 make it ideal for high speed signal processing and video applications.

Operational Amplifier Voltage Booster



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Using the scheme shown above, it is possible to operate most opamps with supply voltages which extend beyond the specified ranges for the opamp on hand. R1-R4 should be chosen to set the quiescent operating voltages on the opamp supply pins to be within the opamp's operating range. In addition to increasing the effective supplies compliance of the op-amp, the circuit also increases the maximum peak-to-peak swing at the output beyond the quiescent supply. Below are the results:

Device	V_{CC}	V_{EE}	V_{out} (max)	t_{rise}	t_{fall}
LM7121	+30	-30	25V	23ns	23ns
LM6171	+30	-30	40V	62ns	46ns

LM7171 - High Output Current Voltage Feedback Amplifier

Features

- Voltage feedback topology
- High Output Current
  - 100 mA into a 100Ω load
- Very High Slew Rate
  - 4100 V/μs, at Vs = ± 15V and Av = +2

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The LM7171 is built on National's advanced VIP™III(Vertically Integrated PNP) complementary bipolar process. The LM7171 is a high speed voltage feedback amplifier that has the slewing characteristics of a current feedback amplifier; yet the LM7171 can be used in all traditional voltage feedback amplifier configurations. The LM7171 is ideal for high speed signal processing such as HDSL and pulse amplifiers due to its high slew rate of 4100V/μs and an unity gain bandwidth of 200MHz. With a high output current of 100mA, the LM7171 can be used for video distribution as a transformer driver.

Applications

Video Distribution

Drive Transformers

Multimedia broadcast systems

LM6311- Low Noise High Speed Voltage Feedback Amplifier

Features

- Voltage Feedback Topology
- Ultra-low Input Noise: 2.3nV/ $\sqrt{Hz}$
- -3dB bandwidth: 110MHz
- Slew Rate: 200V/μs
- Very low Offset Voltage: 0.5mV

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The LM6311's combination of speed, low noise, and distortion and low dc errors will allow the part to be used in high speed signal conditioning applications in order to achieve the highest signal to noise performance. The part is a good choice for high speed applications which require a very wide dynamic range such as the input buffer for a high resolution A to D converter.

Applications:

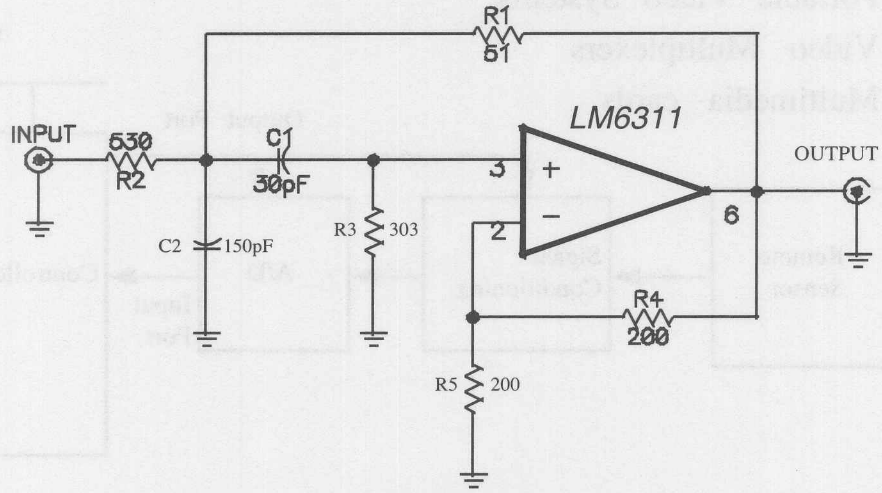
Wide dynamic range amplifiers

ADC/DAC buffer

Pulse/RF amplifier

Low power portable video

2nd Order Sallen-Key Band Pass Filter



The low noise characteristics of the LM6311 make it a good choice for filters.

Circuit shown above has the following characteristics:

Q=20, G (midband gain)=2, f (center frequency)=20MHz

Design Equations:

$C2 = 5*C1$

$G = 1 + R4/R5$

$R2 = 2*Q/[G*C2*2*\pi*f]$ f = Center Frequency

$R1 = G*R2*\{SQRT[1 + 4.8Q^2 - 2G + G^2] + 1\}/[4.8Q^2 - 2G + G^2]$

$R3 = 5*G*R2*\{SQRT[1 + 4.8Q^2 - 2G + G^2] + (G-1)\}/4Q^2$

Procedure:

1- Choose values for R4 and R5 (I.e.. R4=R5=200 to get G=2).

2- Choose reasonable values for C1 and C2 (C1=30pF, and C2=5C1=150pF).

3- Compute R1 and R3.

Note:

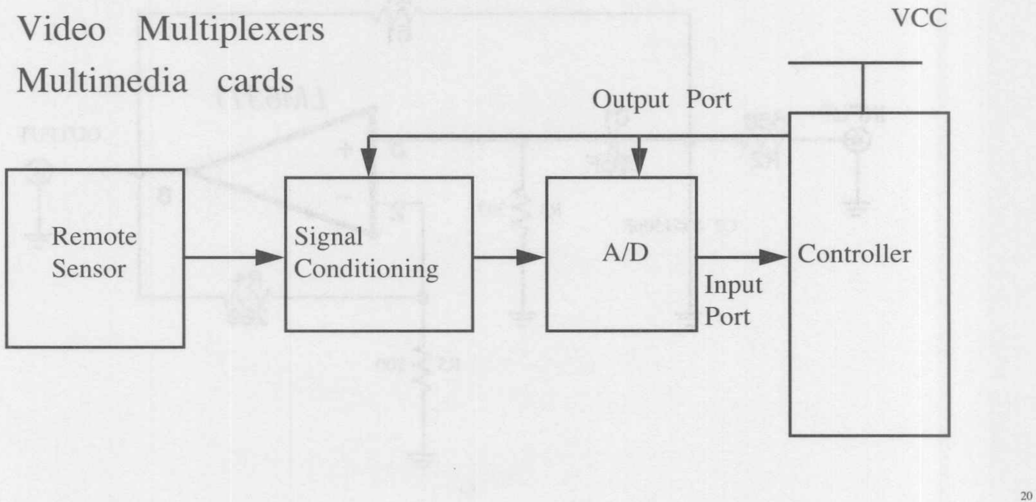
For good high frequency performance, keep the values in the following range:

1k < R < 10k

10pF < C

Power the Analog Portion of a System with a Controller

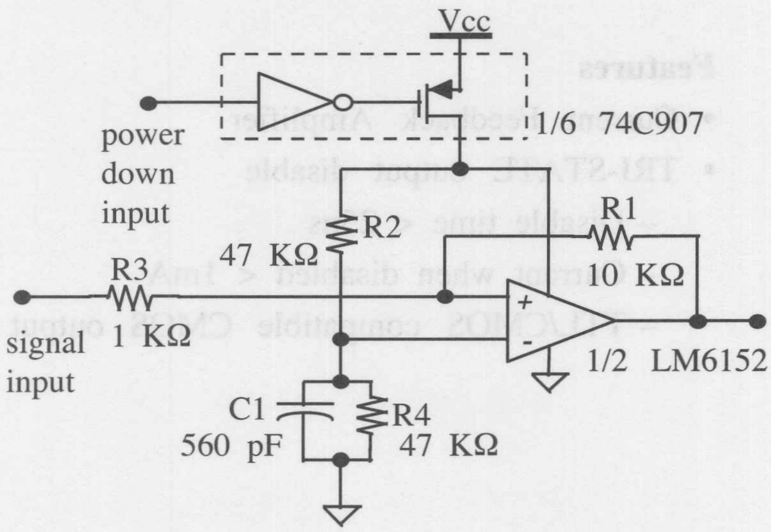
- Portable Video Systems
- Video Multiplexers
- Multimedia cards



As the drive towards portable products continues, systems are rapidly changing to single supply 3V and 5V systems. Analog IC manufacturers are responding to this challenge of low voltage design by introducing devices which operate from lower supply voltages, have lower power dissipation, and smaller footprints. Several techniques can be employed to reduce the power consumption of the integrated devices, and consequently of systems. The above circuit examines a popular method of decreasing power consumption.

The analog portion of the circuit above consists of the signal conditioning block, the remote sensor, and the A/D converter. The analog ICs can be powered on or off by the output port of the intelligent controller.

Power Down Circuit



Using the circuit shown above, it is possible to put the LM6152 into a Power Down or sleep mode, and thereby reduce the power consumption considerably.

The 74C907 is a 6 channel low power open drain buffer used as a switch to control the flow of current to whatever is connected to its output. The output of the device is of the P-channel FET type and is the ideal choice for high side switching. In addition, the 74C907 accepts CMOS compatible inputs which simplifies the interface. Employing high side switching as opposed to opening the ground pin of the device you wish to force into Power Down, has the advantage that depending on the application, in the latter case the supply current might find other paths to ground and complete Power Down may not occur. It is important to tie the inputs to unused channels to ground or Vcc. It is possible to parallel two or more channels of the 74C907 in order to get higher output current capability. At room temperature and 5V operation, each channel can supply 2mA of current with less than 0.4V of voltage drop. With all 6 channels in parallel the voltage drop reduces to 0.4V/6 (or about 70mV). In the Power Down state, the current drawn from a 5V source would reduce to about 16uA.

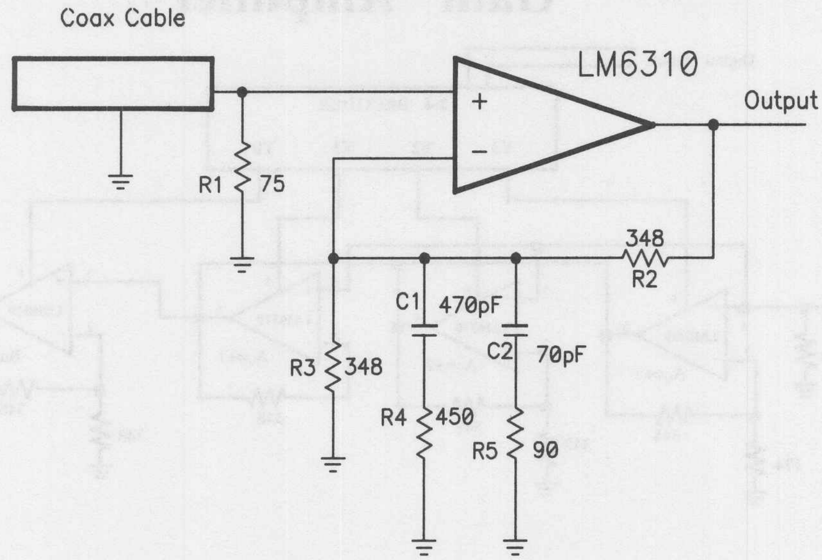
LM6310 - Low Power Amplifier with TRI-STATE Output

Features

- Current Feedback Amplifier
- TRI-STATE output disable
  - – Disable time < 25ns
  - – Current when disabled < 1mA
  - – TTL/CMOS compatible CMOS output

The LM6310 is a video speed operational amplifier with a disable function. The disable function puts the amplifier output into a high impedance state. This makes the LM6310 ideal for distributed video multiplexing. This can enhance manufacturing flexibility by making it easy to add or delete options to basic designs. The disable function can also be used for half-duplex communication.

Amplitude Equalizer



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Sending high frequency signals through coaxial lines of 50 meters or more involves several challenges. One of these is the excessive attenuation of the high frequency components of the original signal. This presents itself as a rounding off of the sharp edges in the waveform similar to what you would observe when looking at a RC time constant limited circuit. What is needed is an amplifier which has progressively higher gain at higher frequencies in order to compensate for this loss in the high frequency region.

Referring to the schematic above, C1 & R4 produce a zero in the transfer characteristics given by :

$f_{z1} = [1/\pi C_1 (2R_4+R_3)]$

$= 540\text{kHz}$

However, this boost in gain diminishes as the frequency approaches the pole produced by C_1 and R_4.

$f_{p1} = [1/(2\pi R_4 C_1)]$

$= 750\text{kHz}$

C_2 and R_5 introduce another zero given by:

$f_{z2} = [1/\pi C_2(2R_5+R_{eq})]$; where $R_{eq} = R_4 \| R_3 = 196\Omega$

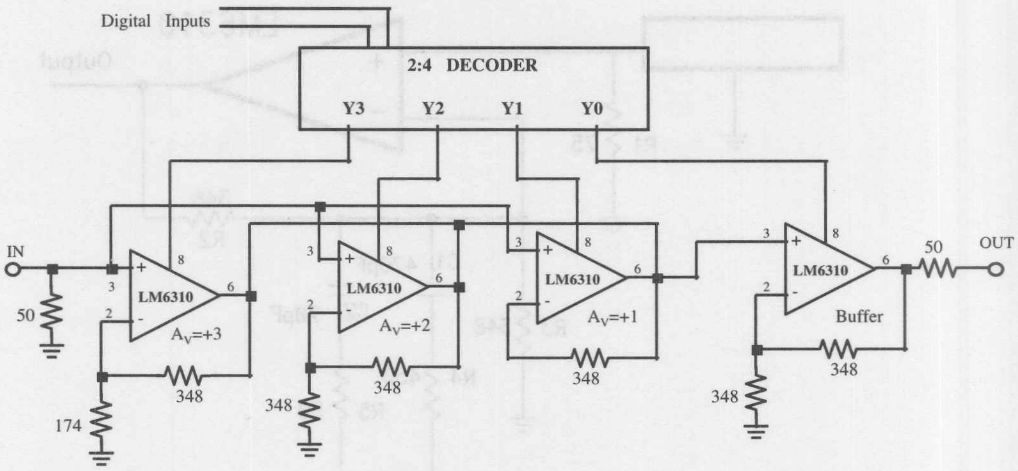
$= 12\text{MHz}$

As in the previous case, this zero will also diminish as the frequency approaches its pole which is given by:

$f_{p2} = [1/(2\pi R_5 C_2)]$

$= 25\text{MHz}$

Digitally Controlled Programmable Gain Amplifier



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The circuit above allows the user to choose four different gains for Vin; a gain of +1, +2, +3, or a TRI-STATE (0V). For example, if Y3 = 1, and Y0= 0, the input signal is multiplied by a gain of +1. Similarly, if Y2 = 1, and Y0= 0, the gain of the application circuit is programmed to be +2. The LM6310 has a power consumption is 8mW, and a very high output impedance when disabled. The LM6310 thus is an ideal choice for applications which are sensitive to power consumption.

LM6132/34, LM6142/44, LM6152/54 SERIES AMPLIFIERS

	LM6132	LM6142	LM6152	Units
Speed	14	25	30	V/µs
GBW	10	17	75	MHz
Power	360	650	1400	µA/Amplifier
>R-R Input	0.25V>V_S	0.25V>V_S	0.25V>V_S	
R-R Output	0.007-4.992	0.007-4.992	0.007-4.992	V
Wide Supply Range	2.7-24	1.8-24	2.7-24	V

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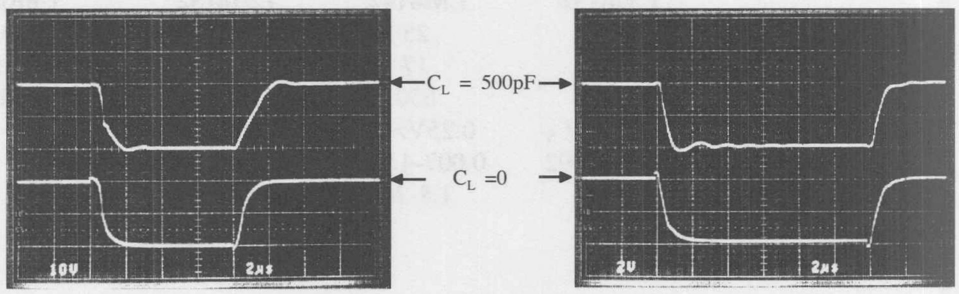
These amplifiers provide world class performance in a complete set of specifications not available in any other amplifiers.

The above specifications plus excellent capacitive loading driving ability make these amplifiers outstanding for many applications, particularly where power consumption is important.

LM6132 DRIVING CAP LOAD



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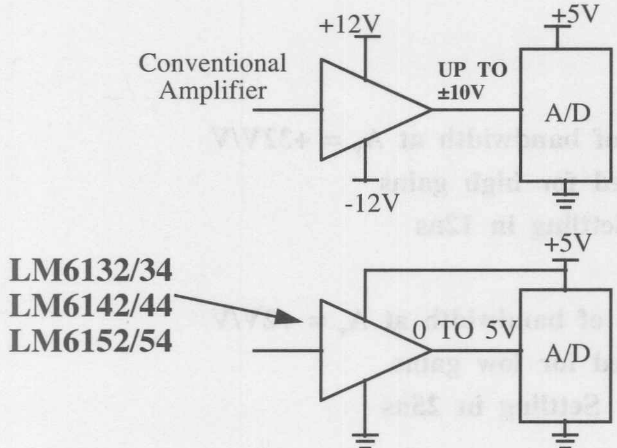
Notice that the rise and fall times remain nearly constant in spite of the greatly increased signal amplitude.

Note also that the settling is only slightly changed in spite of the large capacitive load being added.

Driving A/D Converters with Single Power Supplies



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Many A to D converters have serious problems when their inputs are driven beyond the supply voltages. This is a real possibility when they are driven with a conventional opamp that is operating off of higher voltage supplies to obtain full 0 to 5 volt output. Using an opamp with true, full rail-to-rail output provides the full swing while guaranteeing that 0 and 5 volts will never be exceeded. This can save space and weight, and most of all, save the money that would be spent on the additional power supplies.

CLC501 and CLC502
High Speed Output Clamping Amplifiers

- **CLC501**
  - **75MHz of bandwidth at A_v = +32V/V**
  - **Optimized for high gains**
  - **0.05% Settling in 12ns**
- **CLC502**
  - **150MHz of bandwidth at A_v = +2V/V**
  - **Optimized for low gains**
  - **0.0025% Settling in 25ns**

The CLC501 and CLC502 are high speed amplifiers offering bandwidths of greater than 75MHz and fast output clamps. The CLC501 is optimized for high gains and the CLC502 for low gains:

CLC501 - +7 to +50, -1 to -50V/V

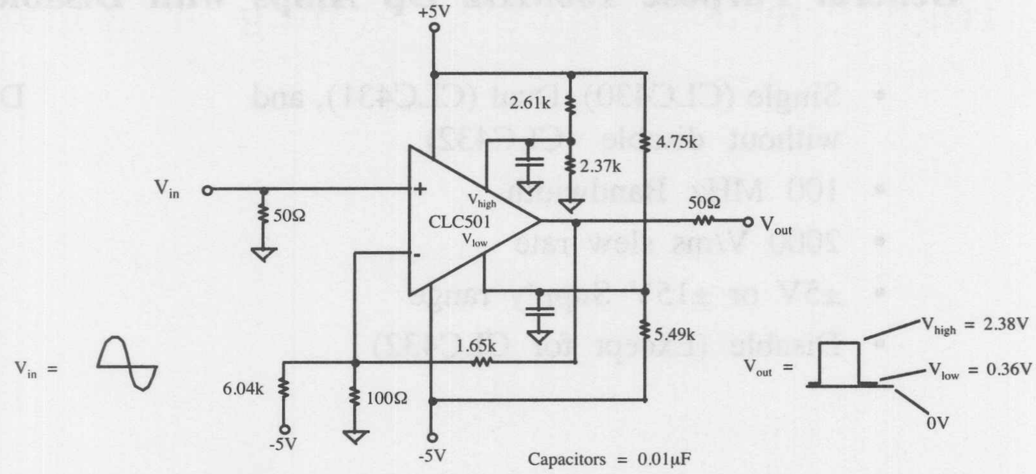
CLC502 - ±1 to ±8V/V

Possible Applications:

Limiting Amplifier driving an A/D

Pulse Amplitude Modulation

CLC501
Low-level AC-to-TTL Converter



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This example utilizes the output clamps of the CLC501. The input sine wave is gained up, level shifted, and clamped to emit TTL output levels.

Circuit Description:

The CLC501 is set for a gain of 17.5V/V. The clamps are set at V_{high}=2.38V and V_{low} = 0.36V.

CLC430/431/432

General Purpose 100MHz Op Amps with Disable

- Single (CLC430), Dual (CLC431), and Dual without disable (CLC432)
- 100 MHz Bandwidth
- 2000 V/ms slew rate
- ±5V or ±15V Supply range
- Disable (Except for CLC432)

30

The CLC430, CLC431, and CLC432 offer high speed, wide bandwidth, and wide supply voltage ranges at a low cost. The CLC430, CLC431, and CLC432 also provide 0.1dB gain flatness to 20MHz (A_v = 2 V/V, $1V_{pp}$)

Possible applications:

Video Distribution

CCD Clock Drivers

Twisted Pair Driver

Switched Gain Amplifier

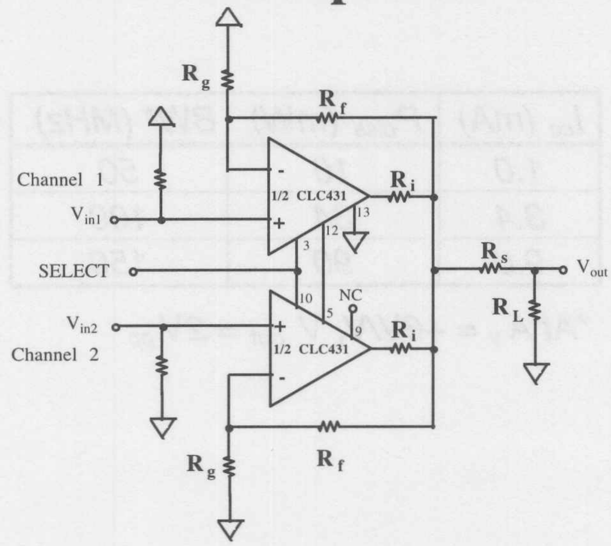
DAC Output Buffers

Imaging Systems

Other high speed amplifiers with a disable feature:

CLC410, CLC411, CLC405, CLC407

CLC431 Application 2:1 Multiplexer



31

This topology utilizes the disable feature and dual package of the CLC431 amplifier. The CLC431 is capable of multiplexing several signals on a single analog output bus. This topology illustrates a 2:1 multiplexer. Implement an N:1 multiplexer in a similar fashion by using an N:1 decoder to enable/disable the appropriate number of CLC431's or CLC430's.

Circuit Description:

The resistor R_i isolates the output of the active channel from the impedance of the inactive channel without affecting the low output impedance of the active channel.

As the circuit is configured above, channel selection is accomplished by applying a logic level to the SELECT node.

Channel 1 - apply logic "low" to SELECT

Channel 2 - apply logic "high" to SELECT

CLC505
Adjustable Power Consumption

I_{cc} (mA)	P_{diss} (mW)	BW* (MHz)
1.0	10	50
3.4	34	100
9.0	90	150

*At $A_v = +6V/V$, $V_{out} = 2V_{pp}$

The CLC505 features a programmable supply current pin. The supply current can be lowered to obtain 10mW power consumption and still maintain 50MHz of bandwidth at a gain of 6V/V. The CLC505 data sheet has a complete set of guaranteed specifications for supply currents of 1mA, 3.4mA and 9mA.

Possible Applications:

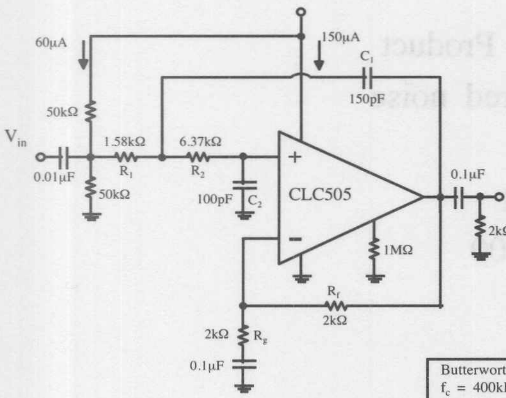
Low Power/Battery Applications

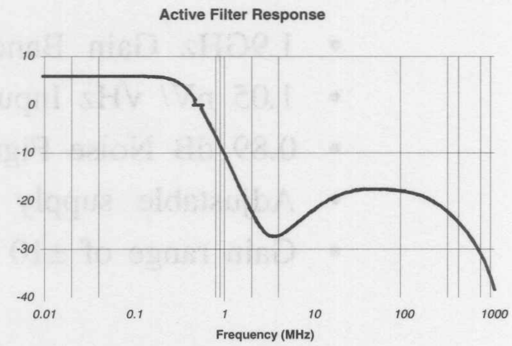
Active Filters

Other high speed amplifiers with adjustable supply current:

CLC425, CLC426

CLC505
Single Supply Active Filter





Butterworth Response
f_c = 400kHz
Amplifier Power = 900µW

33

The CLC505 is well suited for low power filter applications. The circuit illustrated above is a 2nd order Sallen-Key Low Pass filter. The filter was designed for a gain of 2V/V and a corner frequency of 400kHz. The CLC505 is configured in a normal single-supply topology with the input AC coupled.

A simulated response is illustrated above. Comlinear offers spice models for monolithic amplifiers. The models provide excellent results because the input and output stages of the amplifier are kept intact in the model.

Design Equations:

$$Gain = K = 1 + \frac{R_f}{R_g}$$

$$Corner\ Frequency = f_c = \sqrt{\frac{1}{2\pi R_1 R_2 C_1 C_2}}$$

$$Q = \frac{1}{\sqrt{\frac{R_2 C_2}{R_1 C_1}} + \sqrt{\frac{R_1 C_2}{R_2 C_1}} + (1-K)\sqrt{\frac{R_1 C_1}{R_2 C_2}}}$$

For $R_1 = R_2 = R$ and $C_1 = C_2 = C$:

$$f_c = \sqrt{\frac{1}{2\pi RC}}$$

$$Q = \frac{1}{(3-K)}$$

check data sheet for correct info

CLC425
Ultra Low Noise Op-Amp

- 1.9GHz Gain Bandwidth Product
- 1.05 nV/ √Hz Input referred noise
- 0.89 dB Noise Figure
- Adjustable supply current
- Gain range of ±10 to ±1000

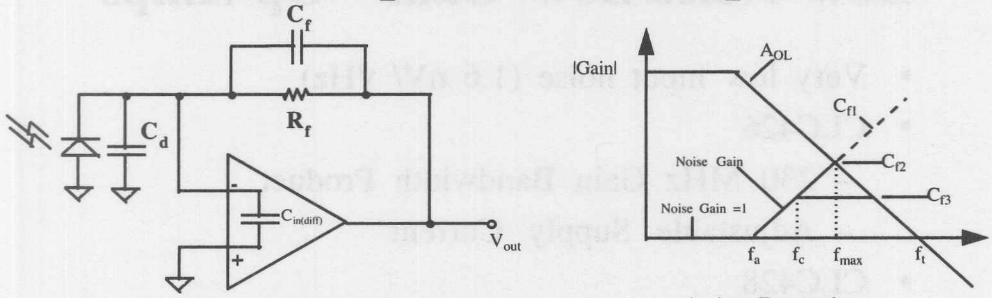
34

The CLC425 is the lowest noise amplifier in its class. The CLC425 not only offers low noise, but also wide bandwidth and adjustable supply current. The CLC425 has been designed into several medical applications as a low noise pre-amp. Another typical application of the CLC425 is a transimpedance amplifier.

Other high speed, low noise amplifiers:

CLC426, CLC428

CLC425
Transimpedance Amplifier



$\frac{Cin(diff) + Cd}{Cf} \geq 10$ for the CLC425 to be stable

At low Frequencies:

Gain = 1

At High Frequencies:

$$Gain = 1 + \frac{(C_d + C_{in})}{C_f}$$

35

For transimpedance applications, a very large gain-bandwidth product is needed to achieve high bandwidths. The CLC425 makes a great transimpedance amplifier because of its 1.9GHz gain-bandwidth product.

Circuit Description:

The desired transimpedance gain is set by R_f. The current times R_f determines the output voltage. The feedback capacitor is needed for stability because transimpedance applications of non-unity gain amplifiers require compensation for the noise power gain at high frequencies. The CLC425 has approximately 1.9pF to 3.0pF for C_{in}. The above is an ideal case scenario.

Graph Definitions:

$$C_t = C_{in} + C_d$$

$$C_{f1} = 0, \phi_m = 0°$$

$$C_{f2} = \sqrt{\frac{C_t}{2\pi R_f f_t}}, \phi_m = 45°$$

$$C_{f3} = 2\sqrt{\frac{C_t}{2\pi R_f f_t}}, \phi_m = 65°$$

$$f_a = \frac{1}{2\pi R_f C_t}$$

$$f_{max} = \sqrt{f_a + f_t}$$

$$f_c = \frac{1}{2}\sqrt{\frac{f_t}{2\pi R_f C_t}}$$

CLC426/CLC428
Low Noise/Low Gain Op-Amps

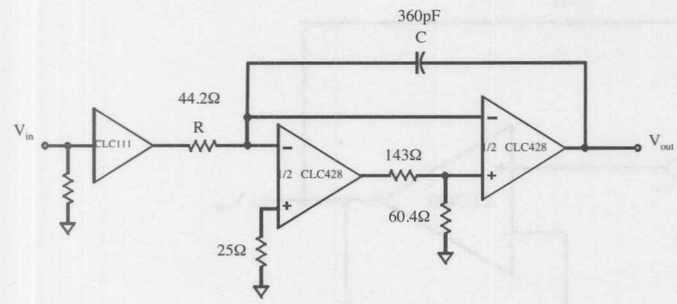
- Very low input noise (1.6 nV/ √Hz)
- CLC426
  - 230 MHz Gain Bandwidth Product
  - Adjustable Supply Current
- CLC428
  - 160 MHz Gain Bandwidth Product
  - Dual Package

The CLC426 and CLC428 are low noise voltage feedback amplifiers. The CLC426 has a minimum stable gain of 2, but can be externally compensated for unity gain. The CLC428 is a unity gain stable dual amplifier.

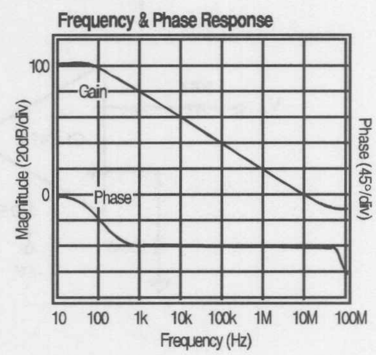
Possible Applications:

Active Filters and Integrators
Ultrasound

CLC428 Composite Integrator



$$V_{out} = \frac{1}{RC} \int V_{in}\, dt$$

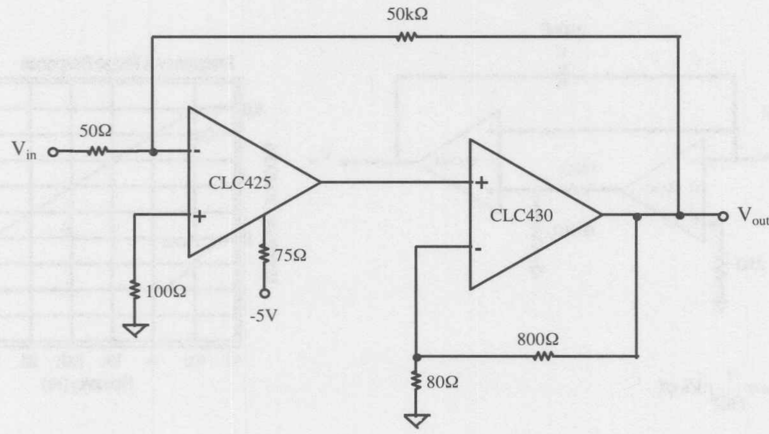


This composite integrator utilizes the dual package of the CLC428 to increase the operational frequency range of the circuit. As indicated by the graph, this topology provides integration over five decades of operation.

Circuit Description:

The 143Ω and 60.4Ω resistive divider reduces the loop-gain and stabilizes the network. R and C set the integrator's gain. The input is buffered by a CLC111.

Composite Circuit



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The CLC425 is a VFB amplifier with a gain-bandwidth product of 1.9GHz. The CLC425 exhibits less than 2MHz of bandwidth at a gain of 1000 V/V. Also, the output swing of the CLC425 is less that ±3.8V. The CLC430 is a CFB amplifier with 75MHz of bandwidth, ±13V output swings, and the gain-bandwidth independence that CFB amplifiers are known for. CFB amplifiers are not well suited for very high gains and are not known for low noise.

Standing alone, these two products could not achieve $20V_{pp}$ output, DC to 10MHz bandwidth, and a gain of 1000V/V. But in the configuration illustrated, this application becomes feasible.

Circuit Description:

The CLC425 is set for a gain of 1000V/V and the CLC430 is set inside the loop with a gain of 10V/V. The CLC425 actually provides a gain of 100V/V allowing it to maintain over 10MHz of bandwidth.

At a gain of 10V/V, the CLC430 uses small resistor values. The noise contribution of those resistors is not significant.

The supply current of the CLC425 is reduced via the R_p resistor to compensate for the phase lag introduced by the CLC430.

CLC407/CLC417
Programmable gain buffers with disable

- CLC407 (Single), CLC417 (Dual, no disable)
- Similar to ‘405 & ‘416 but with internal resistors
- Resistor matching allows precision gains of -1,+1 and +2
- Low Cost

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The CLC407 and CLC417 eliminate the need for external resistors while providing performance to meet high volume applications in the following areas:

Desktop Video

Flash A/D driver

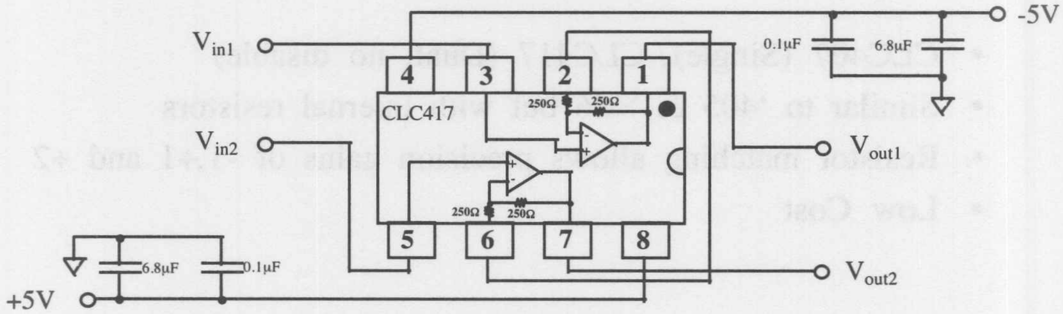
High Source Impedance applications

Video Distribution

Instrumentation Amplifiers

Professional Video

CLC417
Differential Input/Differential Output Amplifier



$$V_{out1} - V_{out2} = (V_{in1} - V_{in2}) \times 2$$

40

In this configuration, the CLC417 provides a differential input/differential output with no external resistors.

CLC440
Very Fast Unity Gain Stable Op-Amp

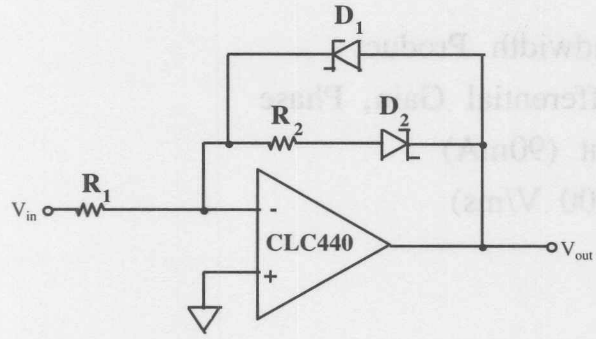
- 230MHz Gain Bandwidth Product
- 0.015%, 0.025° Differential Gain, Phase
- High Output current (90mA)
- Fast Slew Rate (1500 V/ms)

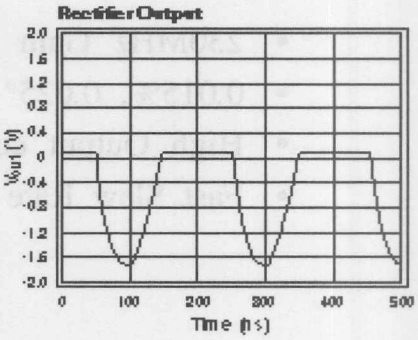
The CLC440 is a voltage feedback amplifier that offers high speed, large bandwidths, and excellent differential gain and phase specs. The CLC440 also offers low noise, 3.5nV/√Hz.

Possible Applications:

Professional Video
Communications
Medical Imaging
Transimpedance amplifier

CLC440
Half Wave Rectifier





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The large bandwidth of the CLC440 allows for high speed rectification. The output for a 5MHz, $2V_{pp}$ sinusoidal input is illustrated.

Circuit Description:

R_1 and R_2 set the gain of the rectifier. Zener diodes are used because of their quicker speed.

CLC449

Ultra-Wideband Amplifier

- 1.2GHz small-signal bandwidth (A_v = +2V/V)
- 2500V/ms slew rate
- 0.03%, 0.02° Differential Gain/Phase
- 3rd order intercept, 30dBm @ 70MHz
- 2.5dB noise figure

43

The DC to 1.2GHz bandwidth of the CLC449 is suitable for many IF and RF applications. In addition, the 0.03%, 0.02° differential gain/phase performance allows system flexibility for handling NTSC and PAL signals.

Possible Applications:

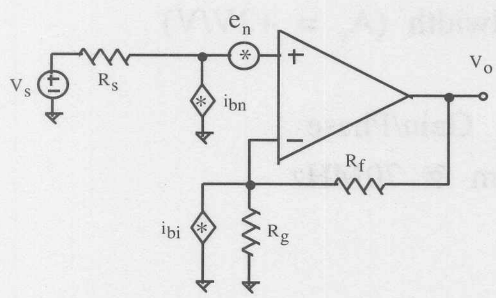
High performance RGB video

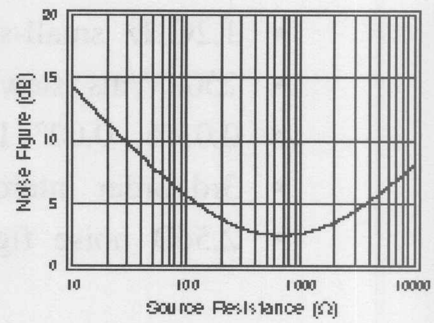
RF/IF amplifier

Instrumentation

Active Filters

CLC449
Noise Performance





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In RF applications, noise is frequently specified as noise figure. This plot illustrates the noise figure of the CLC449 set for a gain of 10 with a feedback resistor, R_f of 100Ω. The minimum noise figure, in this case is 2.5dB and occurs when the source resistance, R_s equals 700Ω.

Variable Gain Amplifiers CLC520/522

- CLC520 - logarithmic gain response
- CLC522 - linearized gain response
- Over 300 MHz Bandwidth Signal Channel
- Over 100 MHz Gain Control Channel
- 40dB effective gain control range
- Differential or Single-ended Input
- Voltage in and out

45

The CLC522 variable gain amplifier (VGA) is dc-coupled with differential voltage inputs and a single-ended voltage output. The maximum gain of the CLC522 is set with two external resistors. Over 40dB of gain range is achieved by sweeping a high impedance voltage input, V_g The nominal maximum gain range is 2V/V to 100V/V. Set at a maximum gain of 10V/V, the CLC522 provides 165MHz signal channel bandwidth and 165MHz gain control bandwidth.

The main difference between the CLC520 and the CLC522 is the gain response. The CLC520 has a logarithmic gain response and the CLC522 has a linear gain response.

Possible Applications:

Variable attenuators

Pulse amplitude equalizers

Video production switching

AGC loops

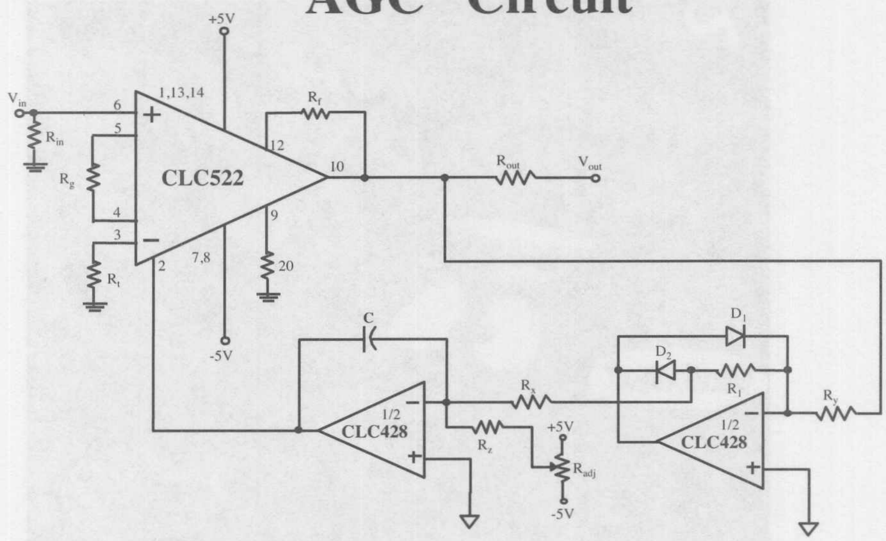
Variable Gain Amplifiers cont.
CLC523

- Low power: 120mW
- 270MHz, -3dB bandwidth
- Gain Range - 2 to 100V/V
- Gain control range - 80dB
- s curve control feature
- 8-pin package

46

The CLC523 is the newest VGA. The low power and convenient 8-pin package are its greatest features.

CLC522 Application AGC Circuit

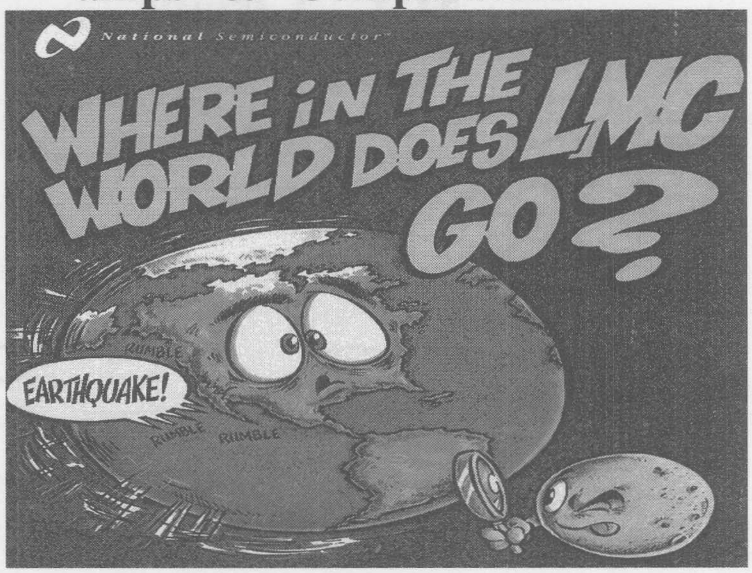


An Automatic Gain Control (AGC) circuit is a typical application for Variable Gain Amplifiers. A typical AGC circuit consists of a Variable Gain Amplifier and feedback loop that performs integration and rectification.

Circuit Description:

The dual voltage feedback amplifier (CLC428) is used to drive the gain control pin of the CLC522. R_1 and R_y set the gain of the rectifier. R_x R_y and C provide a time constant that sets acquire and hold times. The adjustable resistor, Radj, sets the inverting pin of the integrator to the initial condition of +1V. When the RMS current of the signal is greater than the negative current of R_{adj} the integrator decreases the gain of the CLC522. And when the signal drops below the R_{adj} current, the CLC522's gain is increased.

National Semiconductor's CMOS Op amps & Comparators



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LMC stands for Linear Monolithic CMOS and is the prefix of most National Semiconductor's CMOS op amps and comparators; the other prefix is LPC (Low Power CMOS).

Where does "LMC" go?...

CMOS amplifiers are particularly well suited for portable (battery operated) electronics because of their low current consumption and their low supply voltage capabilities. However, they find their way into a broad range of applications because of their relatively broad range of performance characteristics with categories of ultra low power, precision, and respectably high Gain Bandwidth products. You will find National Semiconductor's "LMC" parts in following applications:

Battery Chargers	Helicopters
Antilock Braking Systems (ABS)	Roadside Callboxes
Air Bag systems	Barricade Lighting
Garage Door Openers	Global Positioning Satellites/Receivers
Blood Analyzers	Robotics
Patient Monitoring Systems	Night Vision Goggles

...to mention just a few.

Speed-Power Tradeoff

- LMC6462/4
  - *Micropower Amplifiers (Is =20µA, GBW = 50kHz)*
- LMC6582/4
  - *Medium Power Amplifier (Is =700µA, GBW = 1.2MHz)*
- LMC6681/2/4
  - *Similar specs to the LMC6582/4, and also offers the powerdown mode*
- LM6132/4, LM6142/4, LM6152/4
  - *LM6132/4 (Is =360µA, GBW = 7MHz)*
  - *LM6142/4 (Is =650µA, GBW = 17MHz)*
  - *LM6152/4 (Is =1500µA, GBW = 75MHz)*

The new amplifiers can be thought of in four groups.

One market segment calls for micropower operational amplifiers where the critical spec is the supply current, and the bandwidth and the slew rate are not critical parameters. The LMC6462/4 and the LMC7111 address these applications.

There is also a need for RRI/O amplifiers with higher speed; GBW around 1MHz, and slew rate around 1V/µs. The LMC6582/4 and the LMC6681/4 address this market need.

Finally, the LM6132/4,LM6142/4, and the LM6152/4 address those applications where a GBW of around 2-4MHz is needed.

Where National Semiconductor Stands Out in CMOS Operational Amplifiers and Comparators

- Rail-to-rail Input Common Mode Voltage Range (CMVR)
- Rail to Rail Output Swing
- Low Input Currents
- The availability in the TinyPak™ (SOT23-5)
- Guaranteed specifications and extensive characterization

50

The rail-to-rail CMVR is possible by using the body effect of the depletion device in the input stage. The body effect changes the V_t (threshold Voltage) of the input MOSFETS. By selecting an appropriate V_t of the transistors, the mode of operation can be optimized for rail-to-rail input. For example, the input differential pair MOSFETS behave as depletion devices near the positive rail, and as enhancement devices near the negative rail. Many of our newer CMOS amplifiers offer CMVR that actually exceeds the supply rails. For example, the LMC6462/4 has a CMVR 250mV above the positive rail, and 100mV below the negative rail.

National Semiconductor's CMOS amplifiers have output swings that come closer to the supply rails than other manufacturers. The output swing of an operational amplifier is limited by the Rds On of the MOSFETs in the output stage, and the load that the amplifier is driving. The "wider" output swing can be attributed to lower channel resistance and a stronger driving stage for the output stage, thus fully turning on the pull-up and pulldown transistors. This also results in better ability to drive heavier loads.

National's CMOS amplifiers have lower input currents primarily because of the topology of the input protection circuitry; the protection diodes have a lower leakage current.

We package some of our CMOS amplifiers with TinyPak™ s (SOT23-5) --one of National's newer trademarks. The SOT23-5 measures 3.05mm long x 3.00mm wide x 1.43mm high and is 28% smaller in footprint than an 8-pin SOIC. The SOT23-5 takes less than half of the real-estate than 8-pin SOICs (0.10 square inch versus 0.21square inches).

Why is Rail-to-Rail CMVR Useful?

- Sense signals near the positive supply rail
- Sense signals near ground
- System ground offsets between different equipment

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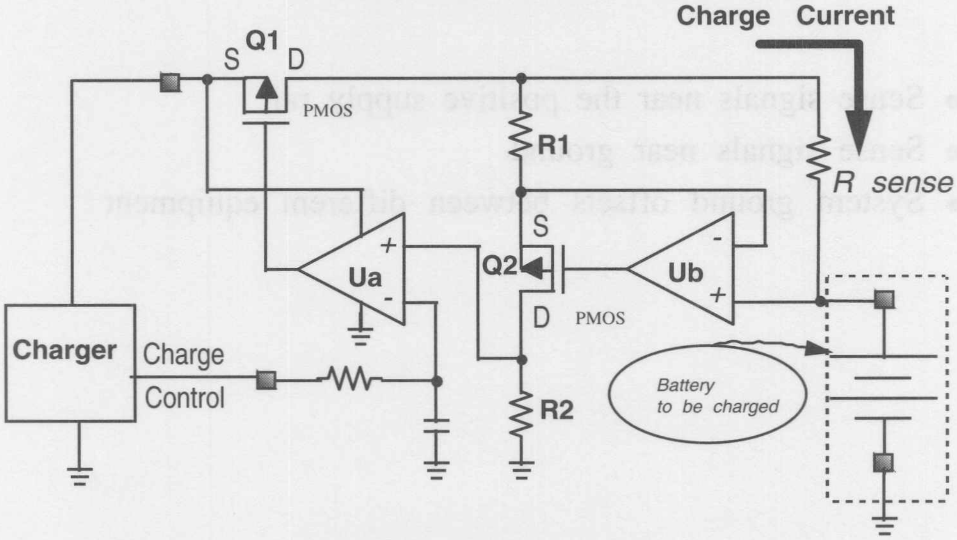
Why is rail-to-rail CMVR useful? Lower voltages in battery powered equipment means that we need to use more of the voltage range. High side sensing, such as measuring the voltage drop across a sense resistor to measure the current charging a battery, requires inputs with CMVR (common mode voltage range) to the positive rail. Ground sensing is necessary with DC sensors that are referenced directly to ground.

Rail-to-rail CMVR can make system design easier by allowing direct interface to a sensor or input signal, and eliminating elaborate bias networks which are required to keep signals in the useful range for non-rail-to- rail op amps.

Finally, when different types of equipment are connected together in a system, their inputs and outputs may shift slightly because of differences in ground voltages. Having a rail-to-rail input reduces the chances of this voltage difference causing system malfunctions.

Current Sensing and Charge Control for Battery Chargers



The above circuit opens or closes the charger's path to the battery depending on the charge current and the charger's control voltage. The charge current flows through the sense resistor which is typically less than 1 ohm. The current in Q2 equals the voltage across Rsense divided by R1. The higher the charge current, the higher the current through R2, which results in higher voltages at the non-inverting input of Ua. Once this voltage becomes larger than the "Control Voltage", Ua turns OFF Q1, thus opening the charger's path to the battery. Hysteresis may be required, depending on the time constants associated with the battery.

To work well in the above application, an op amp needs three characteristics:

1) CMVR to the positive rail.

2) Low voltage offset relative to the smallest current to be sensed by the sense resistor.

3) Good common mode rejection ratio when the inputs are near the positive rail.

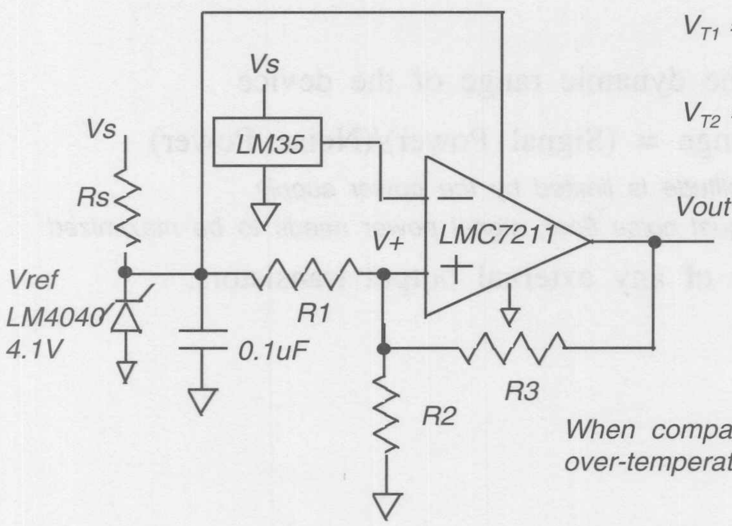
Note that the amplifier runs off the battery being charged or from the charger.

There are several op amps that work in this circuit:

The LM6132. Wide bandwidth with low supply current.

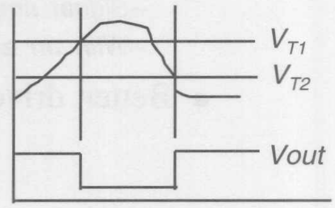
The LMC6482/4 (trimmed Vos) 0.5mV, 500uA/amp.

TEMPERATURE DETECTOR CIRCUIT



$$V_{T1} = \frac{(4.1)\ R2}{R2+R1||R3}$$

$$V_{T2} = \frac{(4.1)\ R2||R3}{R1+R2||R3}$$



When comparator output is low, over-temperature is detected

This circuit is used to detect an over-temperature situation. It can be used in computers to turn on the fan once the temperature inside the computer is too high. The LM35 is used to measure the ambient temperature. Its output voltage increases linearly with increasing temperature, namely, 10mV/degC. LM4040 is used to provide supply voltage to the LMC7211, the comparator. Since the LMC7211 only draws 7uA and operates down to 2.7V, there is no worry using the LM4040 as the power supply of LMC7211

This circuit uses hysteresis to prevent any false triggering if LM35 has a slow changing signal or is operating in a noisy environment. The hysteresis is set by V_{T1} and V_{T2}. From the equations in the slide, one can see that the voltage reference determines the hysteresis. The equations are derived based on the fact that the output of the comparator swings all the way to ground and to positive supply rail.

Why is Rail-to-Rail Output Performance Useful ?

- Maximize the dynamic range of the device
- Dynamic range = (Signal Power)/(Noise Power)
  - *–Signal amplitude is limited by the power supply*
  - *–With an equal noise floor, signal power needs to be maximized*
- Better drive of any external output transistors.

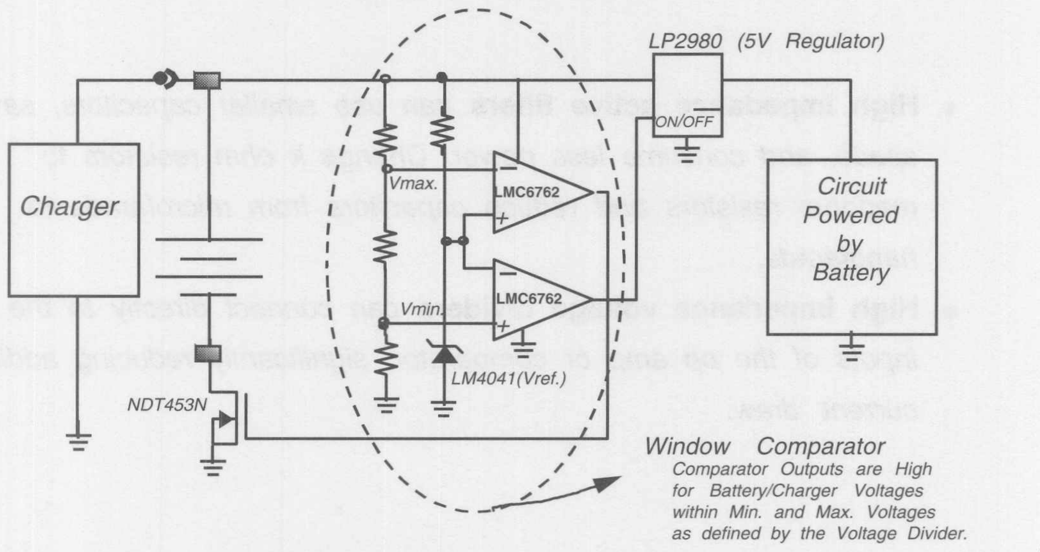
Rail-to-Rail output is useful for a number of reasons...

Lower voltages mean less room for the signal. To maintain the same signal-to-noise ratios, we will need a large signal. This means getting closer to the supply rails.

Rail-to-rail output means a more efficient use of the available supply voltage. This means more power output for a fixed supply. With a rail-to-rail output, you get 277% more power into a load, on a 5V supply, than you will from an op amp that swings to within 1 volt of each supply!

Rail-to-rail output can drive a transistor all the way on or off, which eliminates the need for pull-ups or pull-downs for that purpose.

MIN./MAX. SHUTDOWN CIRCUIT



This circuit can detect a minimum or a maximum voltage of the battery. Once the minimum voltage is detected, the LMC6762 comparator can trigger its output to turn off the LP2980 at its "ON/OFF" control input. The LP2980 then turns off the other circuits and saves the battery from over discharge. Similarly, it can detect maximum voltage of the battery. Once high voltage is detected, the LMC6762 disconnects the battery from the charger by removing the ground connection by switching OFF the NDT453N --a power NMOSFET with ultra low channel resistance. This saves the battery from over charge and the regulator from damage. The LMC6762 is an ultra low power dual comparator with a maximum supply current of 10uA/comparator. It is designed to operate over a wide range of supply voltages: 2.7Vto15V. The LMC6762 has guaranteed specs. at 2.7V to meet the demands of 3V digital systems. The LMC6762 has an input common mode range which exceeds both supplies. This is a significant advantage in low voltage applications. The LMC6762 also features a push-pull output that allows direct connections to logic devices without a pull-up resistor. A quiescent power consumption of 50uW/amplifier (@ V+ =5V) makes the LMC6762 ideal for applications for portable phones and hand-held electronics. Other applications are laptop computers, metering systems, mobile phones, RC timers, window comparators, alarms and monitoring circuits.

Note that the addition of hysteresis may be needed for certain application.

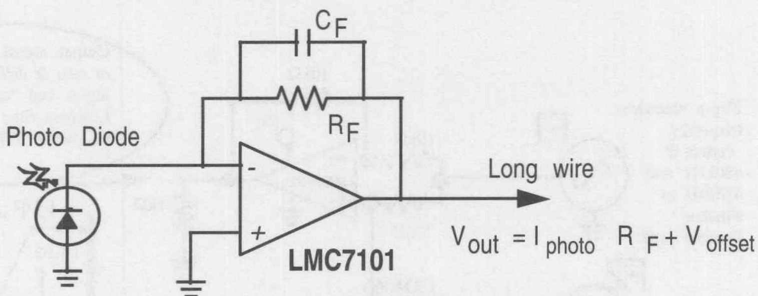
Why are High Impedance and Low Input Current Useful?

- **High impedance active filters** *can use smaller capacitors, saving space, and consume less power. Change k ohm resistors to megohm resistors and reduce capacitors from microfarads to nanofarads.*
- **High impedance voltage dividers** *can connect directly to the inputs of the op amp or comparator, significantly reducing additional current draw.*

Active filters, such as the filter needed for processing the voice signal in a portable phone, consume power. Much of the power used in an active filter is used to drive the reactive elements (primarily capacitors) of the filter. By using a higher characteristic impedance for the filter design, this power can be reduced. The high input impedance of the LMC7101 allows it to be used with a higher characteristic impedance without affecting filter performance. By using separate TinyPak™ amplifiers instead of duals, it may also be possible to improve signal isolation between parts of the active filter.

Voltage dividers at the input of op amps and comparator used for example as to offset an input signal or as a reference voltage can contribute significantly current consumption. CMOS op amps and comparators allow resistors in the megohms to connect to their input; thus reducing power consumption.

Transimpedance Amplifier



Placing the LMC7101 near the photo diode can reduce noise pickup.

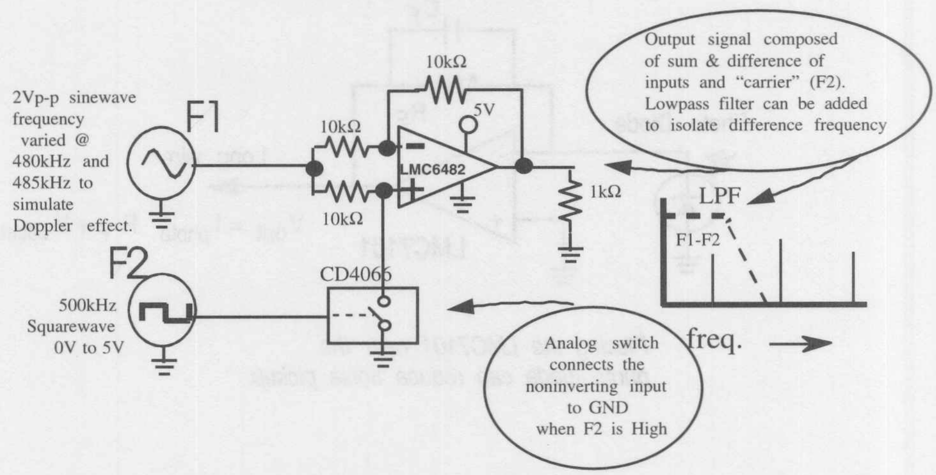
The above circuit illustrates where high input impedance is desirable and maybe even necessary. The photo diode produces only microamps of current. This small current range requires that the op amp's feedback resistor be in the megohm range to produce practical output voltages. With such values for the feedback resistance and for the source current, the op amp's input can't rob current from the photo diode or from the feedback resistor. The LMC7101 op amp meets these needs for the above circuit and more.

Let's look at how we can use the small size of the LMC7101 to improve systems.

In this example, we place the LMC7101 very close to the sensor, (the photodiode) and run a long wire from the output. This long wire is driven by the low impedance of the amplifier output, which will tend to absorb most noise. If we had run a long wire directly from the photodiode to a quad op amp, the wire would be at a high impedance and would be susceptible to noise. The small size of the TinyPak™ lets us improve the system performance by placing each amplifier near the signal source.

This noise can also be reduced by electrical shielding or differential sensing of the signal. By using a small amplifier next to the signal source, it is possible to get more immunity to electrical noise. The LMC7101 thus reduces the size, weight, and cost while achieving greater noise rejection.

SIMPLE MIXER SUBSTITUTE



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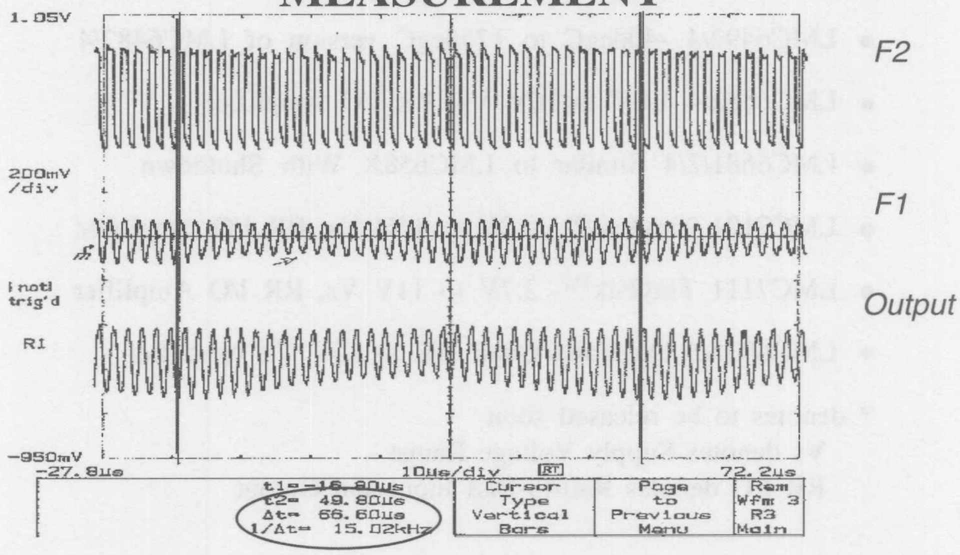
This simple solution provides the ability to extract the difference frequency from two input frequencies (F1 & F2) --a useful function for circuits that utilize the the Doppler effect to measure velocity. F1 could be the reflected or echoed signal that results from F2 hitting a moving object. The reflected signal (F1) enters the simple mixer circuit while the circuit alternates from inverting to noninverting relative to the transmitting signal (F2); F2 switches the 4066 (an analog switch) at the op amp's noninverting input, switching it to ground at the rate of F2. While F2 switches the op amp, it is also transmitted out towards the object for a velocity measurement. The reflected signal F1 is higher or lower in frequency depending on the direction the object is moving in --lower if it is moving away from the receiver, higher if it is moving towards the receiver. The velocity of the moving object directly relates to the difference frequency --the larger the difference frequency, the faster the velocity. A low pass filter can be used to extract the difference frequency from amongst the other frequency components generated by the mixing action of the circuit.

This circuit was built and tested using half of a LMC6482 powered by single 5 v supply that was only required to source 550 uAmps with F1 at 2Vpp, 485kHz, sinewave and F2 as a 5 V, 500kHz, squarewave. (See actual measurements on the following three pages.)

The LMC6482 makes a good candidate for the Simple Mixer because it provides for rail to rail input and output which gives the designer more latitude with signal levels and allows for single supply operation. It also consumes only 500uAmps/Amplifier (typ.) Yet, it provides a GBW of 1.5MHz allowing for a reasonably broad range of operating frequencies to fine tune this circuit for different parameters of the Doppler effect like velocity range and resolution.

THE SIMPLE MIXER'S OSCILLOSCOPE MEASUREMENT



This plot illustrates the the display of the oscilloscope (Tektronix DSA 602A) as it measures the simple mixer's two input frequencies and its resulting output signal. F1 is the middle trace at 485kHz. F2 is the top trace at 500kHz. The bottom trace is the output of the simple mixer. Note that the output frequency contains a 15kHz component which appears as a amplitude modulating wave on the bottom half of the signal. This appeared on the top half also, when the input (F1) was offset above ground. Also note that input F1 did not appear modulated until it was connected to the circuit.

National's Newer CMOS Op amps

- LMC6462/4 20uA/channel, Dual and Quad, RR I/O Amplifier
- LMC6492/4 -40degC to 125degC version of LMC6482/4
- LMC6582/4 1.8V to 10V Vs, RR I/O Amplifier
- LMC6681/2/4 Similar to LMC658X With Shutdown
- LMC7101 TinyPak™ 3V to 15V Vs, RR I/O Amplifier
- LMC7111 TinyPak™ 2.7V to 11V Vs, RR I/O Amplifier
- ⋆ LMC6035/6 Dual and Quad, 3V to 15V Vs Amplifier

* denotes to be released soon
Vs denotes Supply Voltage Range
RR I/O denotes Rail-to-Rail Input and Output 60

National's Newer CMOS Comparators

- LMC6762 10uA/channel, Dual RR I/O Push-Pull Comparator
- LMC6772 10uA/channel, Dual RR I/O Open Drain Comparator
- LMC7211 TinyPak™ 7uA, 2.7V to 15V Vs, Push-Pull Comparator
- LMC7221 TinyPak™ 7uA, 2.7V to 15V Vs, Open Drain Comparator
- ★ LMC7215 TinyPak™ 0.7uA, 2V to 8V Vs, Push-Pull Comparator
- ★ LMC7225 TinyPak™ 0.7uA, 2V to 8V Vs, Open Drain Comparator

* denotes to be released soon
Vs denotes Supply Voltage Range
RR I/O denotes Rail-to-Rail Input and Output

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This slide and the previous slide show National's new CMOS amplifiers and comparators that have been introduced within the last year (CY95). National CMOS amplifiers are well-known for their rail-to-rail input and output feature. In our CMOS amplifier datasheets one can find input common mode voltage range (CMVR) and common mode rejection ratio (CMRR) specified clearly to give you the competitive advantage of good performance over the entire CMVR. Our SOT23-5 is industry's first and only introduction to this tiny package to save precious board space in portable devices. Typical circuit layouts using TinyPak™ take half the space of SO-8.

SPICE UPDATE

- New SPICE Macromodel for the LM6172 Dual High Speed, Low Power, Low Distortion, Voltage Feedback Amplifier Available in Macromodel Library Revision 4.3, April 1996.

- National Semiconductor SPICE Macromodel Library is Now On The WEB @ http://www-analog-mktg/nscspice/nscspice.htm

WWW.natsemi
national

Intelligent Battery Products

Topics to be Discussed

1. What is an Intelligent Battery?
2. Fuel Gauge and Charge Control
3. Envireonmental Battery Characteristics
4. Discharge Profiles
5. Charging Methods
6. Charge Control Flow Diagram
7. Charge Termination Paradigms
8. Fuel Gauge Flow Diagram
9. LMC6980 Product Features
10. Power Management Modes
11. Why a EEPROM is needed
12. LMC6980 ADC Schematic
13. LMC6984/88 Product Features
14. Battery Pack to Host Communications
15. Intelligent Battery Application Schematics
    - Ni-MH with Charge Control
    - Ni-MH without Charge Control
    - Li-Ion with Charge Control

Intelligent Battery Market Profile

Product Segment	Typical Continual Usage	Battery Run Time	Battery Charge Time	Battery Failure Cost	Battery Smarts Today
Portable Computers	8+ hrs	2-3 hrs	1-2 hrs	High	High
Cellular Phones	8+ hrs	10+ hrs	1 hr	Low	Med.
High End Power Tools	4+ hrs	~1 hr	< 1 hr	Low	Low
Radio Control Cars	3-4 hrs	~1 hr	1-2 hr	Med.	Low
Camcorders	2 hrs	1-2 hrs	1-2 hrs	High	Med.

Notebook computers are *the* major users of intelligent batteries

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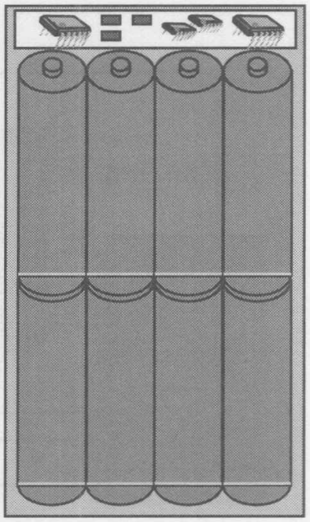
This chart describes the various product markets for battery packs. Only in the cellular phone application does the battery run time exceed the user run time. PC notebooks in particular have a relatively short battery run time of 2-3 hours compared with the desired user run time of 8 hours. The Battery Failure Cost relates to the impact the battery EOD (End-of-Discharge) has on the user. For instance, the Battery Failure Cost is High when the camcorder goes off because the battery is discharged just when an event important to the user is happening.

What is an Intelligent Battery?

- Battery Pack with Integrated Electronics
  - Fuel Gauges the Battery
  - Measures Temperature, Voltage, Current
  - Supports Chemistry Independence
  - Stores Battery Status and History
  - Intelligent Charge Control
  - Provides Battery Protection
  - Communicates with the Host



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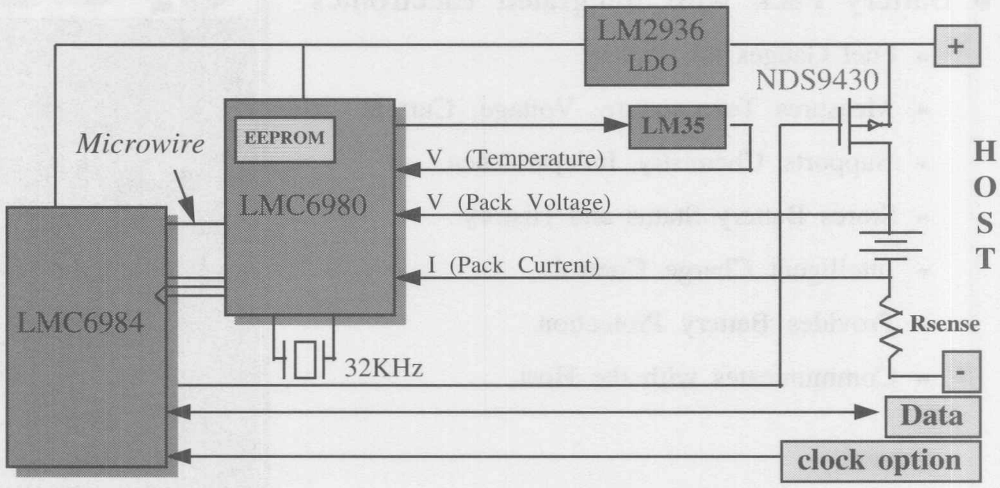
The pictorial is representative of where Intelligent Battery (IB) circuitry can be installed in the battery pack and consume minimum space.

Key reasons to include Intelligent battery electronics in the pack are:

Fuel Gauging the Battery - uses battery current, voltage, temperature and data from the EEPROM correction tables (to be explained later) to compute the battery's State-of-Charge (SOC).

Supports Chemistry Independence - The IB pack looks the same to the host whether Nickel or Lithium based batteries are used. This is because the battery pack performs its own charge control independent of the type of charger used in the Host. The pack also performs its own fuel gauging which is also independent of the battery chemistry. Data in EEPROM tables define the battery parameters to the IB electronics.

LMC6980/84/88 Fuel Gauge & Charge Control Solution



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The LMC6980 in the Intelligent Battery simplified schematic is the data acquisition chip. Its function is to measure the battery's voltage current and temperature, transmit that data over a MICROWIRE serial port so that the LMC6984 can calculate State-of-Charge (SOC). It also has an internal 128 byte EEPROM for storing chemistry specific battery parameters, history data and OEM data. The 32kHz crystal is used to determine the frequency of the system's Real Time Clock.

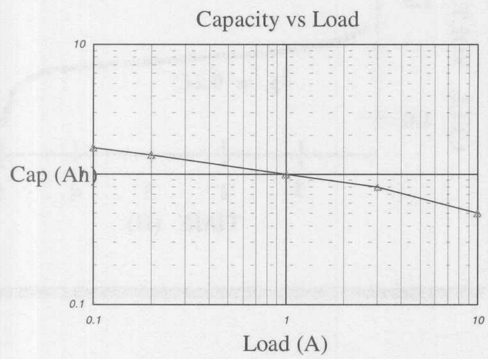
The LMC6984 is an 8 bit COP based ASIC controller, it controls the operation of the LMC6980 based on firmware data stored in its ROM, calculates the SOC, performs power management and battery charge control functions. The LMC6984 is designed for single wire bus applications, the LMC6988 is designed specifically for SMBus applications.

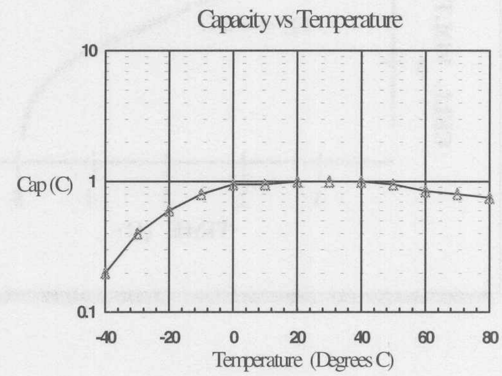
The LM35 is a linear temperature sensor (10mV/°C).

The LM2936 is a micropower LDO regulator (Vin = 40V max., I_D = 7 µA)

Rsense (50mΩ) is used to measure the battery current (50 mV/Amp).

Battery Capacity Vs. Load and Temperature





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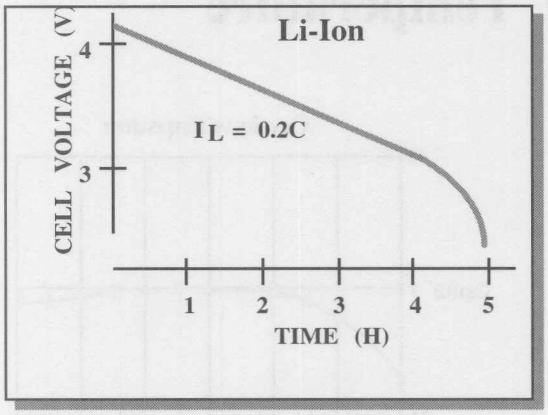
Battery capacity varies as a function of changes in temperature and rate of discharge.

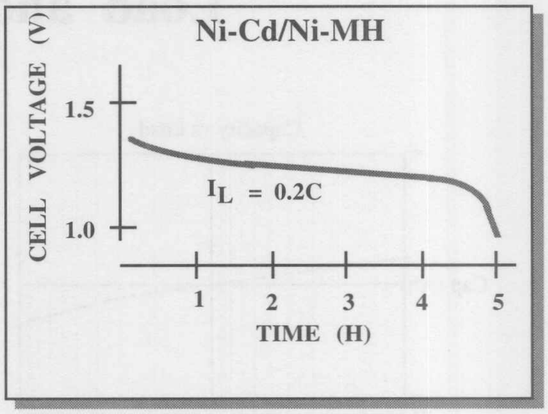
Not shown are graphs for self-discharge rate vs. temperature (rate increases with temperature), charge acceptance vs. charge rate (decreases with increasing rate) and charge acceptance vs. temperature (decreases with increasing temperature).

The details of the graphs are not particularly important, but existance of these environmentally dependent parameters make high accuracy fuel gauging difficult.

These environmental dependent parameters are battery type specific and these data must be stored in the battery pack's EEPROM to achieve high accuracy fuel gauging.

Discharge Profiles





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The higher nominal voltage (about 3.6V) of Lithium is a big advantage over Ni-Cd and Ni-MH because you don't need to stack as many cells in series to get useful voltages.

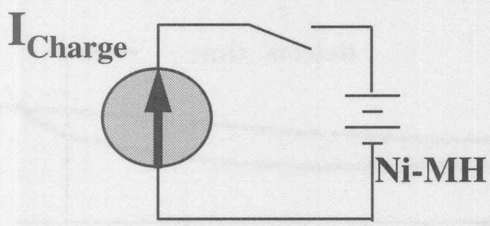
One Li characteristic that is not as good as Ni is the highly sloped discharge voltage. The typical Li-Ion cell will vary from about 4.2V (max.) to about 2.2V (min) over the discharge cycle. A varying discharge voltage is an *advantage* in fuel gauging: an accurate measure of battery state-of-charge can be obtained through a simple voltage measurement, only if the battery is subjected to a constant load during most of its discharge cycle. The high ESR of Li-Ion batteries negates using voltage measurement for fuel gauging when widely varying loads are present, as in notebook computers. However, the non-constant Li voltage requires the design of power converters that operate efficiently over a wide input voltage range, switching converters are the best choice for this application.

As a result, some manufacturers will be forced to learn a new technology (switching regulator design) that they did not previously use. Although this may be an annoyance in the short term, it will eventually yield products with more efficient power supplies.

Charging Methods

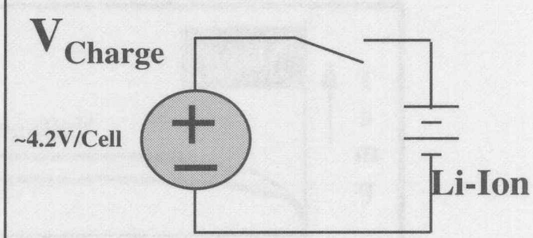
Constant Current (CC)

- Required for Ni-Cd/Ni-MH
- End-of-charge detection required to prevent overcharge



Constant Voltage (CV)

- Used for Li-Ion
- End-of-charge detection not required



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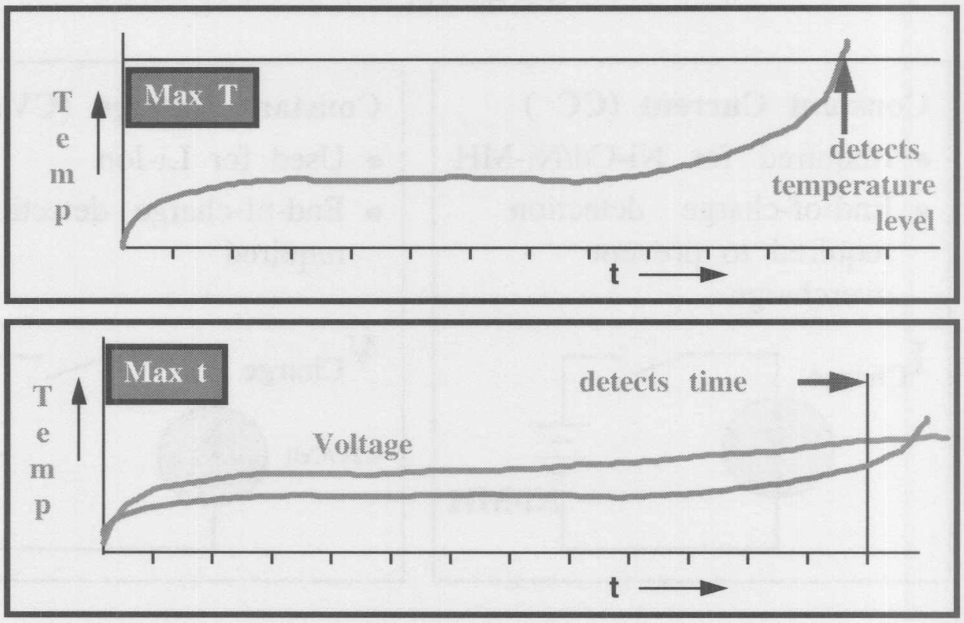
Charging characteristics of batteries are becoming more important as users demand faster recharging. The minimum standard of acceptance for a good consumer product is typically a one hour fast charge (two hours may be OK).

Fast charging of Ni-Cd and Ni-MH require some rather comprehensive (and expensive) charge termination circuitry to avoid overcharge damage.

Two completely different charging methods are used for Ni and Li batteries. A constant-current technique is best for Ni, but a constant-voltage, current limited charger must be used for Li.

The different charging characteristics of the two chemistries is a problem, as most equipment manufacturers would like to do a seamless transition from Ni-MH to Li, using charger that accepts both battery types would be ideal.

Indicators for charge termination



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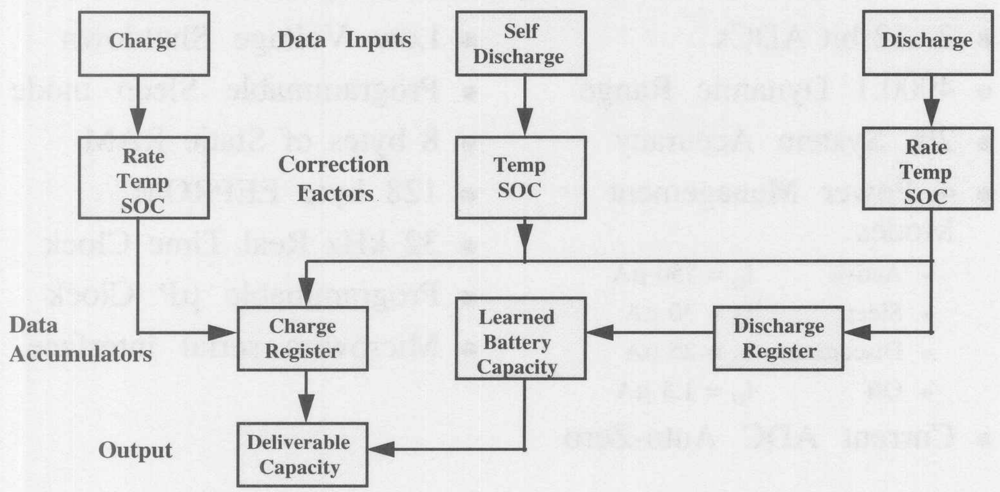
The top graph represents the temperature change in a Ni-MH battery as a function of charge time. This represents the primary method of terminating a 1C charge rate. A typical value for the battery temperature's rate of change is about 1°C/min.

A typical secondary termination method is shown in the lower graph. Here, a maximum period of time is used as a backup for primary method.

These secondary termination methods are are used as safety nets, if for any reason the primary termination limit is not reached before the secondary.

Fuel Gauge Flow Diagram



This flow diagram describes the process by which the remaining or depleted capacity of the battery is calculated.

Battery Charge and Discharge Current data from the ADCs in the LMC6980 is combined with corresponding Rate, Temperature, and SOC EEPROM Correction Factor data.

Self Discharge data is derived by applying Self Discharge vs. temperature and SOC correction data from the EEPROM with elapsed time to derive a correction factor to SOC. EEPROM corrected data for Charge is added to the remaining SOC.

EEPROM corrected data for Discharge is subtracted from the remaining SOC. The IB learns how much the "real" capacity of the battery is by counting coulombs during a complete discharge cycle. This "real" capacity calibration is done when the battery is discharged continuously from 100% charge to 0%. The Learned Battery Capacity factor is used to make a final correction to establish the true Deliverable Capacity of the battery.

LMC6980 Product Features

Battery Data Acquisition System IC

- 2 12 bit ADCs
- 4000:1 Dynamic Range
- 2% System Accuracy
- 4 Power Management Modes
  - Active $I_D = 750\ \mu A$
  - Sleep $I_D = 30\ \mu A$
  - Disconnect $I_D = 25\ \mu A$
  - Off $I_D = 1.5\ \mu A$
- Current ADC Auto-Zero
- Low Voltage Shutdown
- Programmable Sleep mode
- 8 bytes of Static RAM
- 128 byte EEPROM
- 32 kHz Real Time Clock
- Programmable µP Clock
- Microwire serial interface

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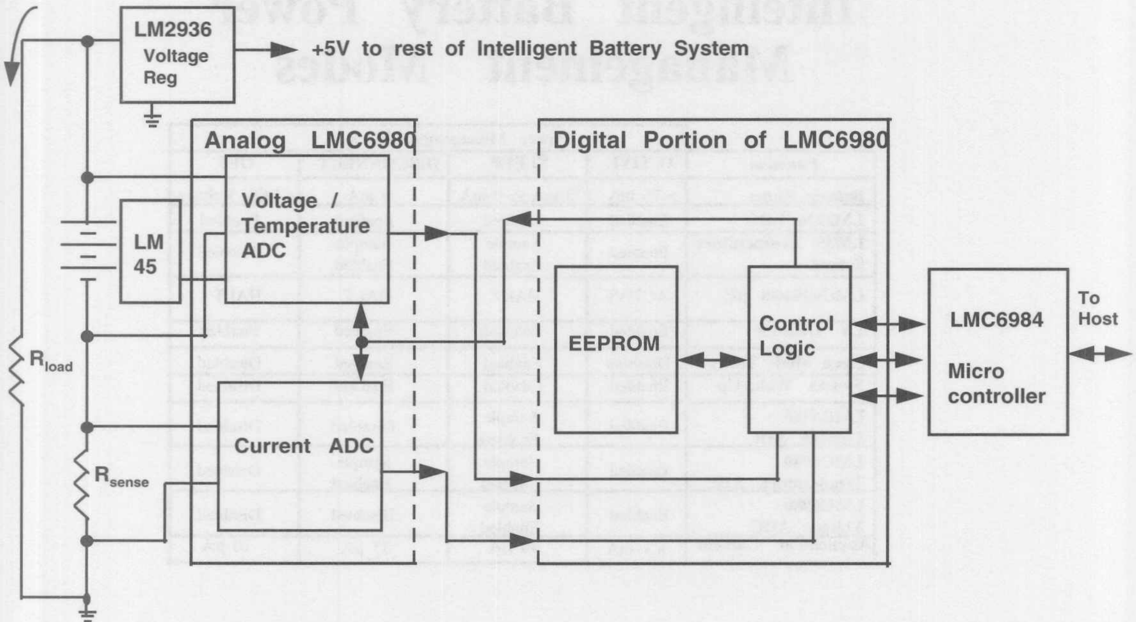
The purpose of the LMC6980 is to acquire the battery's voltage, temperature and current, convert each to a clock synchronized frequency proportional to the voltage, temperature and current then store those data into 16 bit counters. The LMC6984/88 microcontroller retrieves the data from the counters and the EEPROM through the MICROWIRE serial interface. The microcontroller uses these data to calculate the SOC per the flow diagram from the previous slide.

The 3 most important features are:

4000:1 Dynamic range - This wide range is required by notebook computers because of their sophisticated power management systems. Typical current drains range from 1mA in standby mode to 3-4mA when the system is booting. The auto-zero feature insures reliable battery current measurement down to 1mA by effectively reducing the Vos of the current input amplifier to a few mV.

128 byte EEPROM - This feature gives the OEM the flexibility, with one part, to make modifications to the battery correction tables on the fly. The OEM may need to do this if a new battery vendor is used, if an improved battery is replacing an older technology, or if the current battery is replaced by one with a completely different chemistry.

Power Management - The 4 power modes are used so that current drawn from the battery during periods of low or no activity is at a minimum. You don't want the battery to discharge due the IB circuitry while the pack sits on the shelf for 2 weeks.



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This is a very simplified block diagram of the IB's major elements, with the 6980 detailed to show it's major blocks.

Two ADCs are used. One acquires battery current by measuring the voltage across Rsense (typically 50mW) the other acquires battery voltage via a resistor voltage divider (not shown) and temperature using an LM35/45 linear temperature sensor. The LM35/45 has an output voltage of 10mV/°C from 0 °C to 100 °C.

The EEPROM in the LMC6980 block is used to store battery correction data (details in a later slide).

The Control Logic block is multi-purpose. It has control registers to control power management and the analog block's operations, Real Time and Microcontroller clocks, MICROWIRE serial interface, and serial-to-parallel and parallel-to-serial converters to convert the MICROWIRE interface to the 8 bit parallel bus used internally by the LMC6980.

The LM2936 is an LDO 5V regulator used to provide power to the LMC6980, LMC6984 and LM35. Its unique feature is that it automatically shuts itself down to 7mA when its load current goes to zero. This is important, particularly in the OFF mode (described in the next slide) where current drain from the IB circuitry must be at an absolute minimum to prevent the battery cells from forward biasing.

Intelligent Battery Power Management Modes

Function	Power Management Mode ACTIVE	SLEEP	DISCONNECT	OFF
Battery Status	> 75 mA	75 mA to 0 mA	0 mA	< EOL Voltage
LM2936 LDO	Enabled	Enabled	Enabled	Enabled
LM35 Temperature Sensor	Enabled	Sample Enabled	Sample Enabled	Disabled
LMC6984/88 µC	ACTIVE	HALT	HALT	HALT
LV Detector	Enabled	Enabled	Enabled	Enabled
Sleep Mode Timer	Disabled	Enabled	Enabled	Disabled
System Wake-Up	Enabled	Enabled	Enabled	Disabled
LMC6980 Current ADC	Enabled	Sample Enabled	Disabled	Disabled
LMC6980 Temperature ADC	Enabled	Sample Enabled	Sample Enabled	Disabled
LMC6980 Voltage ADC	Enabled	Sample Enabled	Disabled	Disabled
Application Current	8.3 mA	40 µA	35 µA	10 µA

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The chart describes how the various LMC6980 blocks behave in each of the four power management modes.

The Battery Status current is the Host drain current. The 6984/88 firmware provides the actual trip points for each of the power management modes.

In ACTIVE mode, all functions are on 100% of the time.

In SLEEP mode all functions are Sample Enabled (active for 5%, off for 95%, for example) except for those functions required to provide power and wake up the IB circuitry internally, or from the Host.

DISCONNECT mode is the same as SLEEP mode except that current and voltage measurements are not made (the Low Voltage Detector is still on).

In OFF mode all 6980 functions are off and the 6984/88 is HALTed. Just the LM2936 and the Low Voltage Detector are active. Only a voltage greater than the EOL voltage will reactivate the IB circuitry.

The EEPROM Provides Flexibility

- Battery Correction Tables
  - Charge Efficiency vs. Rate, SOC, Temperature
  - Discharge Efficiency vs. Load, Temperature
  - Self Discharge vs. Time and Temperature
- Operating Information
  - Charge Termination Parameters and Limits
  - Exposure History
- Manufacturer's Information
  - Pack Configuration
  - Serial Number and date code

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About 2/3 of the data stored in the EEPROM are for Battery Correction Tables.

Most of the other data is devoted to storing the charge termination limits for the Fast and Slow charge paradigms.

Exposure History data is updated in the EEPROM as needed. Data stored here represents; # of times the battery has properly terminated Phase 1 and 2 charge cycles, what parameter terminated the late charge phase, min. and max. temperature and voltage the battery has been exposed to.

The EEPROM also stores battery pack configuration data related to # of cells, chemistry, and size of cell. The pack's serial #, and the manufacturing date code are also stored.

LMC6984/6988 Product Features

Battery Management System Controller

- 8 bit µProcessor core
- Fully static CMOS design
- 3 Power management modes
  - ACTIVE ID = 7.5 mA
  - IDLE ID = 1.3 mA
  - HALT ID < 1 µA
- 16k bytes ROM
- 256 bytes RAM
- Full duplex UART
- Four Programmable I/O ports
  - Two 8 bit I/O
  - 4 bit Output, 4 bit Input
- Four software selectable I/O options
- Clocked at 10 MHz in most applications
- LMC6984 designed for single wire data transfers
- LMC6988 designed for SMBus applications

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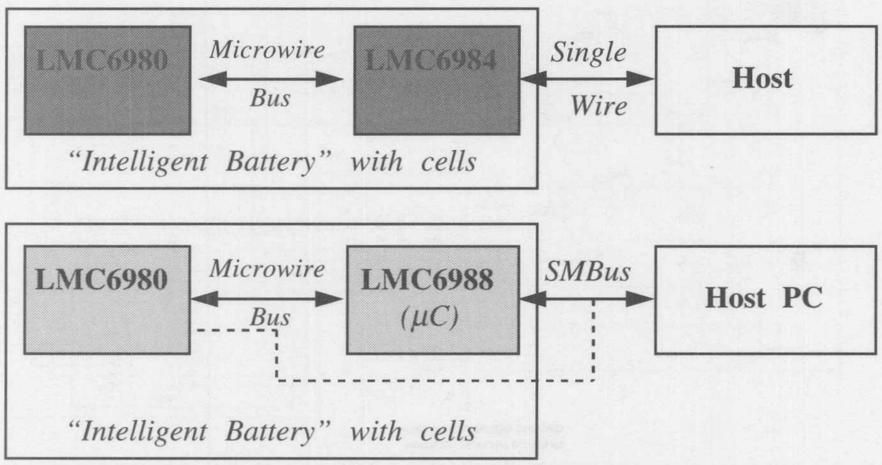
The LMC6984/6988 Controllers have ROM based firmware designed either for single wire communications (LMC6984) or SMBus applications (LMC6988).

The controller has three power management modes, only two of which are used in Intelligent Battery applications (ACTIVE and HALT). There are a total of 24 programmable ports 16 of which are I/O and 8 others with are dedicated inputs or outputs.

The masked 16K ROM is used as permanent memory for the application specific firmware program, the 256 bytes of data memory includes RAM, data registers, I/O registers, and Control registers.

The Full Duplex UART uses the MICROWIRE communications interface as its protocol. MICROWIRE is a four wire serial communications protocol used to transmit and receive data from the LMC6980. The four lines are: DO (data out), DI (data in), SK (data clock), CKI (microcontroller clock).

System Management Bus (SMBus)
in the Intelligent Battery

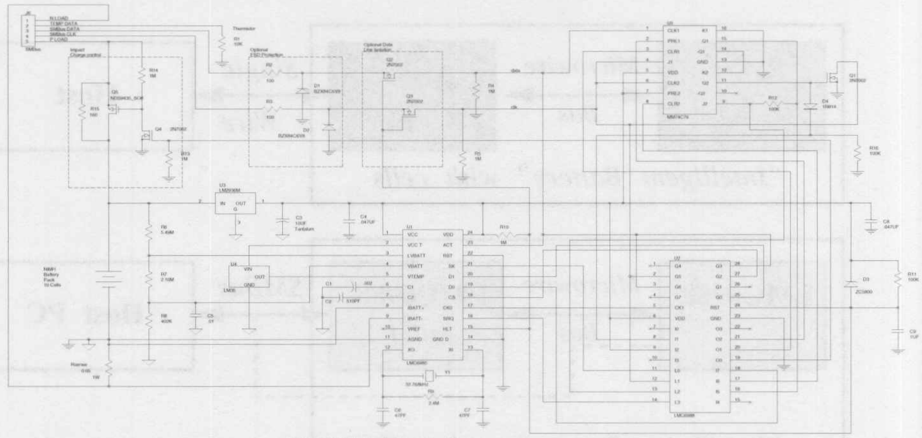


The two diagrams show that the both LMC6984 and 6988 controllers communicate with the LMC6980 through the MIRCOWIRE interface, but communicate with the Host using a completely different interface, determined by the ROM firmware program.

The LMC6988 is designed for use with an SMBus. SMBus is an open drain (or collector) two wire (clock and data) bus sponsored by Intel, with roots derived from the I^2C and ACCESS bus. It is used mainly by slow speed peripherals, such as mice or keyboards, connected to PCs and Notebooks. 128 master or slave devices can be connected to the bus. It transmits data at between 10K and 100K Bits/Sec.

The LMC6984 is designed for use with single wire interfaces, The single wire interface usually has one master and multiple slaves. Most of the time, the IB will act as a slave with the Host acting as master. The bus uses an open drain architecture, similar to the SMBus, but has a maximum data rate slower than SMBus. Bus data is read and written by using time slots to manipulate the bits, and a command word to specify the transaction.

SMBus Compliant Intelligent Battery Application with Charge Control



LMC6980 SMB Application Circuit
for Level 0 and Level 1 Chargers

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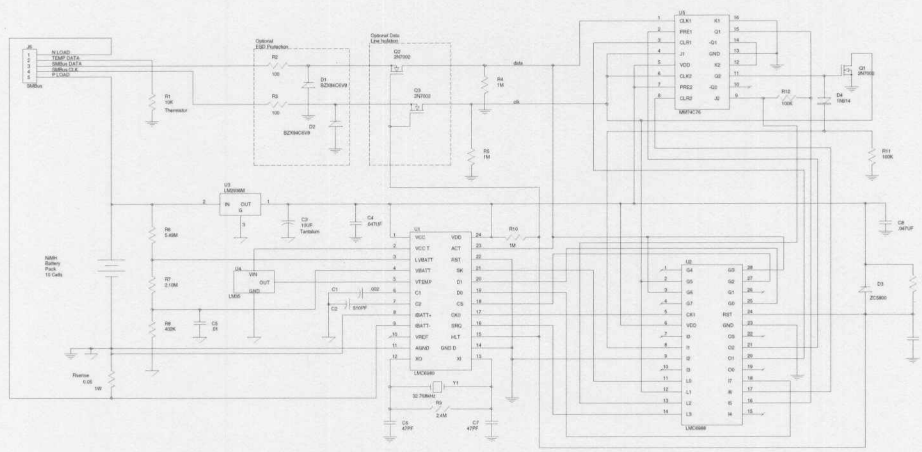
Finally, we get to the nitty-gritty stuff. Real applications. Here and on the next two slides are three Intelligent Battery applications using Ni-MH and Li-Ion batteries with no charge control and in-pack charge control.

As you can see, there is more circuitry than was disclosed in the simplified block diagrams shown previously. The additional components are the 76C74 Dual DF/F (U5) and the 2N7002 (Q1) to support the SMBus. Q2, Q3, R4, and R5 provide line isolation between the SMBus and LMC6988 when the pack is inactive. And R2, R3, D1, and D2 are used for ESD protection. R6, R7, and R8 are used to scale the
battery voltage down to voltage acceptable to the LMC6980 (0V to 1V). The tap connected to LVBATT is used to sense when that voltage falls below $V_{DD}/2$, this voltage is equivalent to the battery's End-of-Life voltage.

The NDS9435 (Q5) is a power PMOS transistor with an $R_{DS(on)}$ of 35mΩ used for controlling the average charge current into the battery. R15 is used to determine the Maintenance Charge current. Q5 is off during the Maintenance Charge phase.

Y1, R9, C6, and C7 form the tank circuit used for the 32.768 kHz Real Time Clock oscillator. The crystal recommended for most SMT applications is the Epson MC-306. 32.768 kHz was selected for its extremely low cost, it's used in millions of quartz watches.

SMBus Compliant Intelligent Battery Application with Charge Control



LMC6980 SMB Application Circuit
for Level 2 Chargers

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C1 and C2 are timing capacitors used for the current and voltage ADCs. The values of C1 and C2 are not critical, as they only determine the peak-peak voltage of the sawtooth waveforms at pins 6 and 7. If they are within 25% of the values indicated in the schematic, the accuracy of the ADCs are not affected.

The schematic on the previous page was designed to used with “dumb” Ni-CD or Ni-MH chargers. These chargers can not be controlled directly and are unaware of the chemistry of the battery they are charging.

The schematic on this slide is identical to the previous slide, except the charge control PMOS transistor, Q5 is removed. This application is designed for use with Ni chargers which perform their own charge control and are aware of the battery’s chemistry.

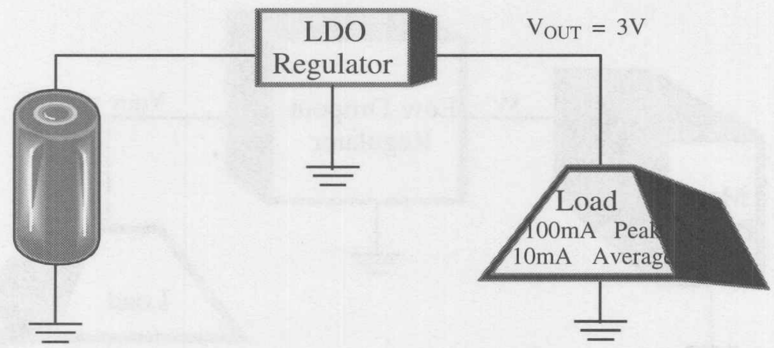
POWER MANAGEMENT

Power Management

- Low Current Power Conversion for Portable Electronics
- Precision Voltage Control
- Li-Ion Cells for High-Density Power
- Power Conversion in Line-Powered Equipment
- New SIMPLE SWITCHERs™ Converters
  - LM2594/5/6/7/8/9 step down buck converters
  - LM2585/6/7/8 flyback/boost converters
  - LM2825 1Amp DC/DC Converter Power IC
- Switchers Made Simple 4.2

2

Low-Dropout Regulator Operates Longer

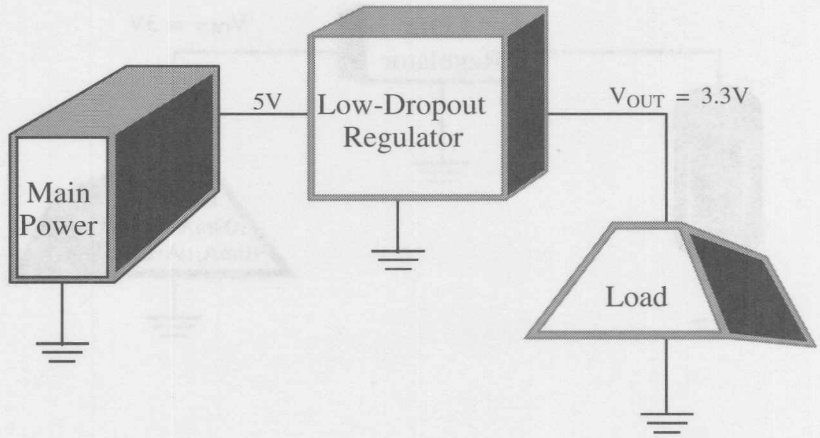


The main power supply will often be subjected to pulses of load current, as different sections become active. It must also have a low-power shutdown mode, which will not drain the batteries at a rate faster than their self-discharge.

Many portable systems, such as a cellular phone, also have a limitation on their physical size. This, in turn, limits the size or height of the power supply. Most components must be surface-mountable, and often the height is limited to less than 1cm.

A low-dropout (LDO) linear regulator is often a good choice for this main regulator, if the load is light and the battery voltage is very close to the desired power supply voltage. The LDO can offer a compact solution with few additional coponents being required. In addition, if the LDO has a tight tolerance for its output voltage, the minimum battery voltage needed to maintain regulation can be accurately predicted.

LDO Power for Peripherals

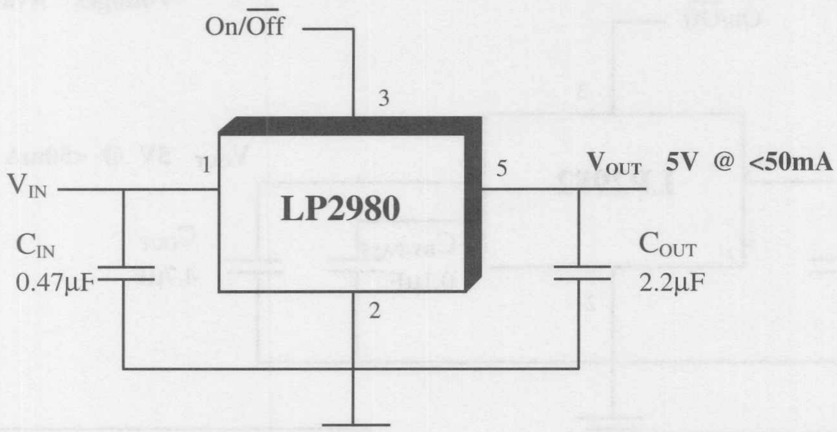


Power for the peripheral parts of the sytem is usually developed from the main supply. In this way, if the main supply is turned off, then the peripherals are also disabled.

The peripheral power conversion still needs to fit in a small area, and meet other restrictions similar to those of the main power supply. However, there may be additional requirements for diagnostic signals. These may include "flags" that indicate whether the converter is properly operating, or which detect other faults within the system.

Low-dropout regulators may also be a good choice for this type of power conversion. Many of the precision LDOs include diagnostic functions that would otherwise need to be added as individual components.

Tiny Bias Supply for Light Loads



Voltages Available
5.0
4.0
3.8
3.6
3.5
3.3
3.2
3.1
3.0
2.9
2.8
2.7

- ± 0.5% Output Tolerance
- <1 µA in Shutdown

To develop a light-load bias supply for a small control section within the portable system, a small linear regulator is the best choice for a small footprint and low power loss. As in this example, a low-dropout linear regulator can provide a 5V, 3.3V, or 3.0V output (depending on version selected) at up to 50 mA, with peak currents of up to 100 mA. With an initial accuracy of +/-0.5% for the A-grade part, the LP2980 also has sufficient precision to be used as a reference.

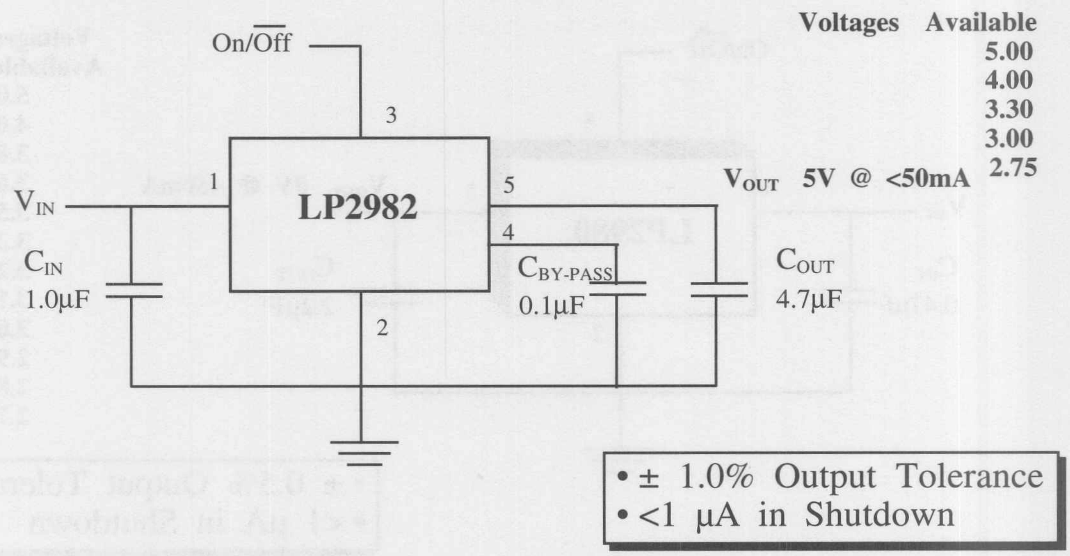
At the maximum steady-state load of 50 mA, the dropout voltage of the LP2980 is only 225 mV (maximum, over temperature). In comparison to the conventional 0.6V LDO dropout, this very low dropout voltage allows the battery to discharge further before the output voltage loses regulation. The LP2980 dropout voltage is also specified at lower current levels, for accurate planning of battery life.

When the regulator is not needed to be operational, it can be put into shutdown mode during which it draws less than 1 µA. With this low quiescent current, it is not necessary to disconnect the regulator with a series FET switch.

As with other low-dropout regulators, the LP2980 requires an output capacitor. The value of this capacitor must be at least 1 µF, over all operating conditions; 2.2 µF is preferred for variation with temperature and tolerance. For most applications, a good surface-mount tantalum capacitor is appropriate for the task.

The LP2980 is available in the SOT23-5 package (only 1.6mm by 2.9mm). Although this small package size minimizes the board area required by the regulator, it also limits the power dissipation (yet staying within the temperature range limits).

LP2982 Low Noise Output

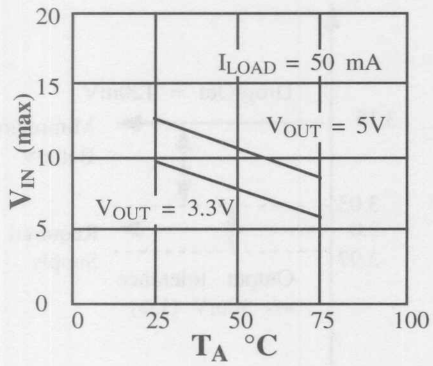


The LP2982 provides a low noise output with the same power as the LM2980. The reference has been connected to pin 4 to allow for addition of a by-pass capacitor. This by-pass capacitor reduces broadband noise and reduces typical noise to 40uV on the output.

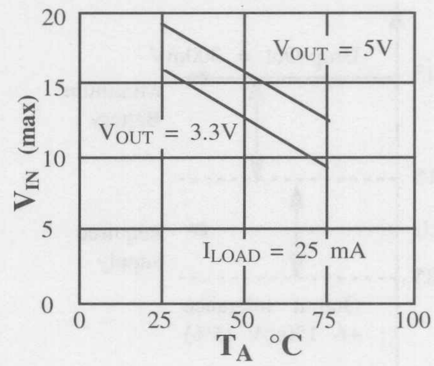
The output capacitor selected must be a high quality ceramic of film capacitor to keep leakage current to a minimum. Leakage currents greater than 10nA can cause excessive load on the reference circuit and reduce accuracy of the output.

Practical Input Voltage Range

SOT23-5 Thermal Resistance Limits LP2980 Max. Input Voltage



Max. input voltage at *full* load



Max. input voltage at *moderate* load

The internal (junction) temperature of the LP2980, given the actual operating conditions, should be calculated before finalizing the regulator design. This junction temperature is the sum of the maximum ambient temperature, and the temperature rise due to the operation of the regulator. The maximum T_J limit of the LP2980 is 125°C:

$$T_J = T_A + \Delta T_J\ ;\quad T_{J(MAX)} < 125^{\circ}C \qquad \text{Eq. 1}$$

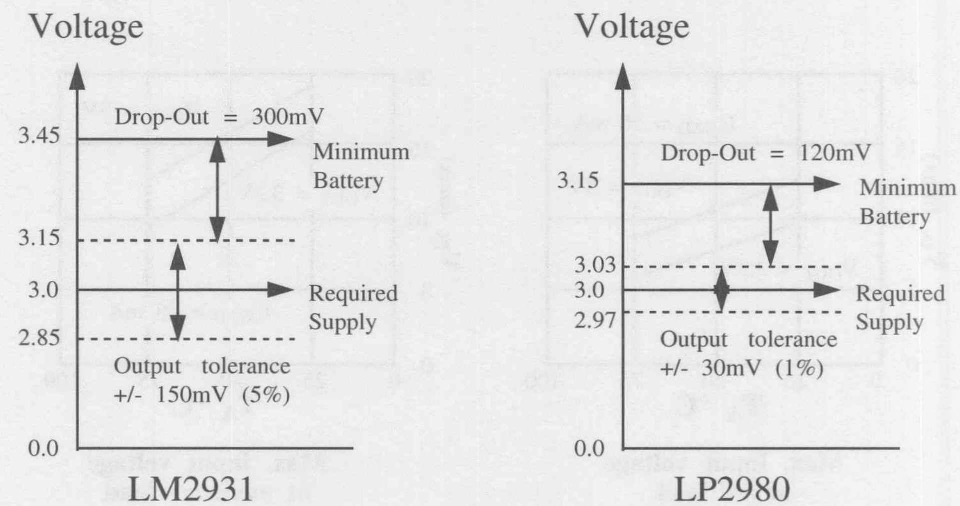
where the temperature rise (ΔT_J) is the product of the device power dissipation and its package thermal resistance:

$$\Delta T_J = P_D \times \theta_{JA} = [(V_{INmax} - V_{OUT}) \times I_{LOAD} + V_{INmax} \times I_S] \times \theta_{JA} \qquad \text{Eq. 2}$$

where V_{INmax} is the maximum voltage that will be applied to the LP2980, I_{LOAD} is the maximum load current in the application, and I_S is the regulator supply current. θ_{JA}, the thermal resistance, is 300 °C/W for the SOT23-5 package.

While the maximum input voltage of the LP2980 is 20V, if it is to be used with the full 50 mA load current at room temperature, the practical limit without exceeding the junction temperature limit is about 11.5V. Operation at lighter loads increases the practical input voltage range; higher ambient temperatures decrease the voltage range.

Typical Drop-Out Characteristcs

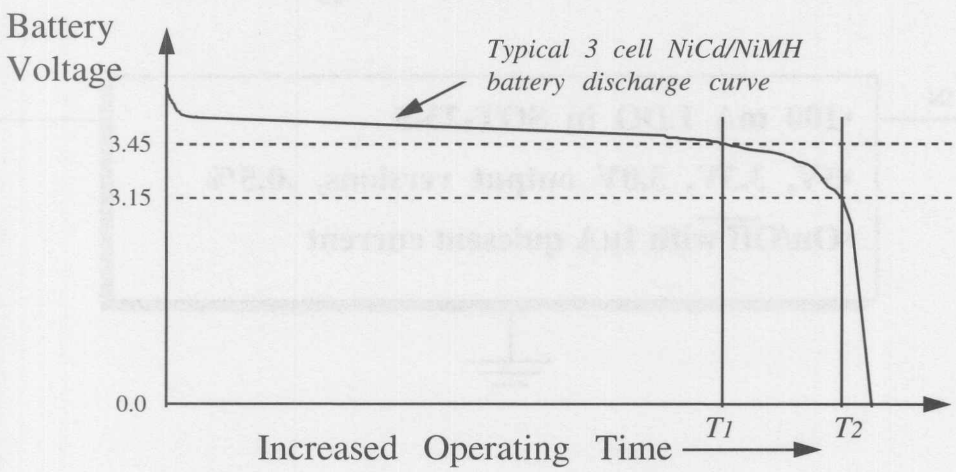


It has already been commented that lower dropout performance enables a battery to discharge further before the LDO starts to encounter regulation problems. The output accuracy of the regulator also has an effect on the lowest operating battery voltage required for regulation, and by optimising these two parameters, a battery powered system is able to achieve a longer run time.

Let us consider a typical battery powered system which needs to supply 50mA at 3V from a battery pack consisting of 3 NiCd or NiMH cells. One solution would be to use an LM2931, one of the first LDO regulators to be produced by National in the 1980s. This part will have a typical accuracy of +/-5%, and a typical dropout of around 300mV. Combining the worst case of maximum output voltage with the drop-out, the LOWEST battery voltage that such a regulator could operate with is 3.45V.

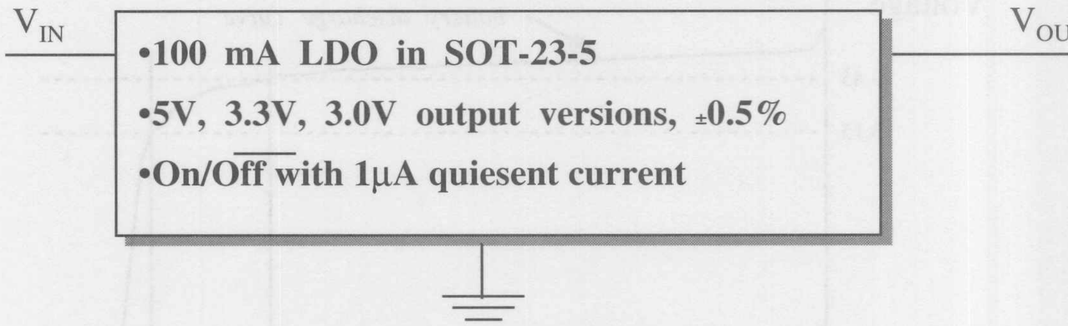
Compare this with the LP2980 which has a typical accuracy of +/-1% and typical drop-out of 120mV. This regulator would still operate with a minimum battery voltage of 3.15V, a 300mV improvement.

Operating from reduced battery voltage extends system run-time



The effect of reducing the minimum operating voltage has an obvious impact on the operating time of the system. The typical discharge curve (shown above) for a pack of 3 NiCd/NiMH cells starts to drop significantly in voltage as the cells become fully discharged. By reducing the minimum battery voltage required for operation, more of the battery packs power can be used and operating times extended. In todays market environment, equipment runtime is a key feature used to differentiate products.

NEW LP2981
Similar Features, Higher Load



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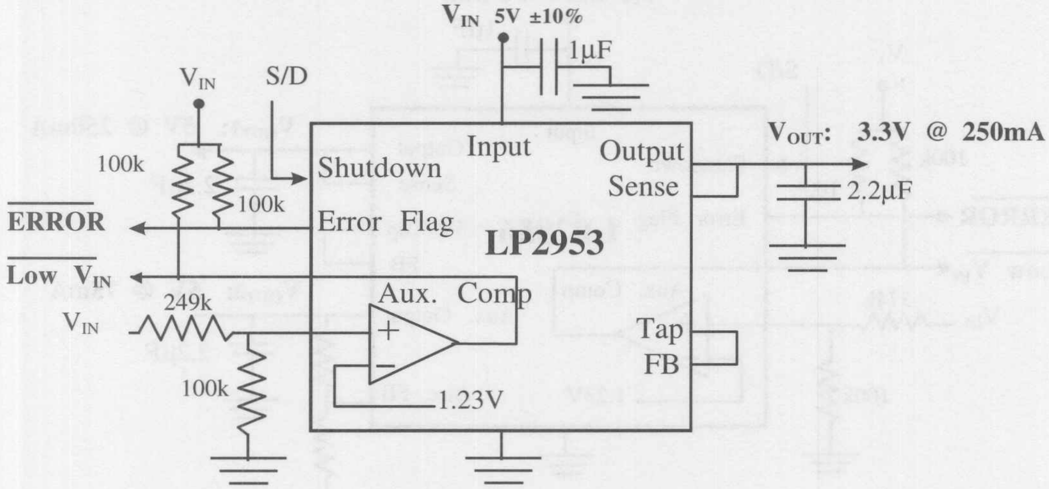
The LP2981 offers the same features as the LM2980. However, greater caution must be exercised when using the full output current of this device due to the thermal limitations discussed earlier. Also, a 4.7µF capacitor is recommended on the output because of the higher average load.

Low-Power Bus Conversion



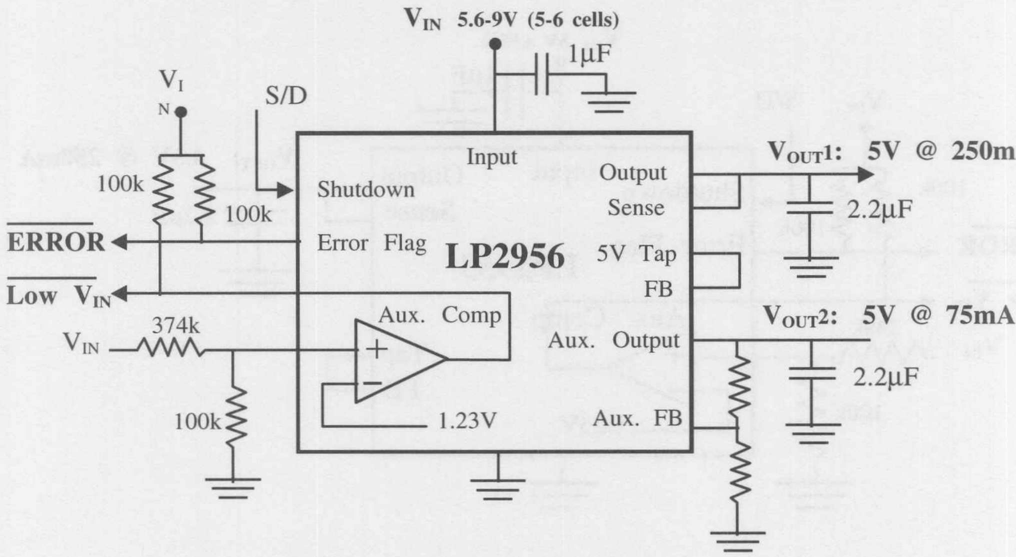
Within a given portable system, there are often components which operate from a +5V supply as well as those which need a +3.3V supply. If the power required by the +3.3V load is relatively low, then the simple approach of developing the 3.3V using a linear regulator may be best.

While the efficiency of a linear regulator can be no better than the ratio of the output voltage to the input voltage (e.g. 66% for a 5V-to-3.3V conversion), the linear regulator may take less board area than a switching regulator for the same output power. In addition, several low-dropout linear regulators are available with special system features to allow them to interface easily with digitally-controlled systems.

There are several regulators that are able to convert a 5V supply to 3.3V. For load currents of up to 250 mA, the 3.3V version of the LP2953 low-dropout regulator can be used. In this application, the LP2953 is 58% efficient at a 1 mA load, only consuming 2.4 mW. With a 100 mA load, the efficiency rises to 63%, with only 0.2W of power loss. When used as a 5V-to-3.3V converter, the LP2953 does not require special heat-sinking for ambient temperatures less than 80°C.

The LP2953 also incorporates a shutdown function, provides an error flag, and has an auxiliary comparator for monitoring other signals within the system. In the example of the above figure, the auxiliary comparator is monitoring the input voltage, and will send a logic-low signal if the input voltage drops below 4.3V. This signal indicates that the 3.3V output is close to losing regulation, and the system should either begin to shut itself down, or otherwise prepare for possible misoperation.

Dual Regulated Outputs



Within many systems, it is often necessary to have different regulated outputs powering different parts of the system. This improves fault tolerance, so that a fault within one section will not affect the operation of the rest of the system. This is especially useful when there is memory that must have uninterruptable power at all times.

The LP2956 low-dropout regulator is intended for this type of use, with two independent, fully-protected regulated outputs. The main output is rated for up to 250 mA, while the auxiliary output is rated for 75 mA. If the main output is shut down, whether by current limit or a shutdown command, the auxiliary output remains operational.

In addition, diagnostic features in the LP2956, such as the auxiliary comparator and the Error Flag, are available to indicate the regulator's status to a microcontroller. The auxiliary comparator can monitor the input voltage, and detect when it falls below a threshold (determined by the resistor values in an external divider). In this example, this low-input threshold is set to 5.8V. This signal indicates that the regulator is close to dropping out of regulation.

If the input voltage sags low enough, the main output drops out of regulation (i.e. drops more than 5% below the nominal value). At this point the Error Flag, an open-collector output, goes low. Both open-collector outputs should be pulled up to either V_{OUT2}, or to the system logic supply, so they remain valid even when V_{OUT1} is shut down.

Precision 0.25A Low-Dropout Regulators

Part #	Diagnostics	Other Features
Dual In-Line & Surface-Mount		
LP2952	Shutdown, Error Flag	
LP2953	Shutdown, Error Flag	Comparator, Precision Ref.
LP2956	Shutdown, Error Flag	75mA Aux. Reg., Comparator
TO-220 & TO-263 (Cropped Tab Surface Mount)		
LP2954		
LP2957	Shutdown, Error Flag	

Tolerance for all parts is (A) ± 1.4%; Tolerance is maximum, over temperature, line, and load ranges.

There are several other 250 mA LDO regulators that are similar to the LP2953 and LP2956. These include the LP2952 and LP2957 single regulators (without comparator), and the LP2954 3-terminal regulator.

The maximum dropout voltage or each of these regulators is 600mV at full load (800mV over temperature). The dropout voltage is also specified at lighter loads.

For the regulators with shutdown control, the regulator is put into a low-power mode by driving the SHUTDOWN pin to a logic-low level. In this mode, the output is turned off, but the regulator control section is left operational so that the LDO can quickly regain regulation (usually within 0.5ms, with an output capacitance of 2.2 μF). In shutdown mode, the supply current is typically 105 μA.

Each of these LDOs is available with an output voltage tolerance of ±1.4% (maximum) over temperature, line, and load ranges. The operating temperature range is -40°C to +125°C

Similar Features, Higher Load Current With LP2960

V_{IN} V_{OUT}

- **500 mA LDO in 16-pin SO, DIP**
- **5V, 3.3V output versions, ±1.4% tolerance**
  - **System features**
  - **Error Flag**
  - **Auxiliary Comparator, precision reference**

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When the peak load current is above 250 mA, the 500 mA regulator LP2960 should be considered. It has the same high accuracy (±1.4%) as the LP2953 and LP2956, and also offers the features of error flag and auxiliary comparator. The key difference is in the higher output current range.

The LP2960 output voltage can be set to the internally-programmed value of either 3.3V or 5V (depending on version selected). If a different voltage is needed, it can be set to any level between 1.23V and (V_{IN} -1V). The maximum input voltage is 30V.

New Automotive Voltage Regulators

When Safe Operation is "Not An Option"

Safety Critical Systems

Safety Airbags

Anti-Lock Brakes

Medical Equipment

Security Systems

Avionics

These Regulators ...

Interference Energy Sources

Electro-magnetic Interference (EMI)
Power Lines

Radio Frequency Interference (RFI)
Transmission Towers
Mobile Radios

Bulk Current Injection (BCI)
Wiring Harness Pick-up

Supply Line Transients
Automotive Load Dump

Remain in Regulation ! **Do Not Generate A System Reset Flag !**

Fail-Safe and Fault-Tolerant systems are becoming more prevalent throughout the industry. A major source of interference is high frequency energy which can bombard electronic modules from a variety of sources. As an example, in automotive a major concern is the energy radiated by high powered police radios which at any time can be in close proximity to a vehicle equipped with an anti-lock braking system and a safety airbag. This high level of RF energy cannot interfere with braking operation or, worse yet, inadvertently deploy an airbag. In cases such as these safe operation is *not an option*, it's a *given*!

The voltage regulator is a key component in any system and must remain in normal operation or the system fails immediately. These three new voltage regulators are the beginning of a new family of products where design techniques have been employed to provide immunity to high frequency interference. By immunity we mean that these products have been tested and proven to remain in normal regulation and do not generate a false system reset while exposed to energy levels ranging from 1MHz to 400MHz with a signal strength of up to 300V/m.

Without the voltage regulator in full operation, no system would have a chance for continuous functionality.

Three New Low-Dropout *System* Voltage Regulators

Part Number	Output Voltage	Output Current	Features
LM9070 **Single**	**5V**	**250mA**	**HF Tolerant** **Keep-Alive Logic Control** **Delayed Reset Flag**
LM9071 **Single**	**5V**	**250mA**	**HF Tolerant** **Delayed Reset Flag**
LM9072 **Dual - Tracking**	**5V Main** **5V Tracking (1%)**	**350mA** **80mA**	**HF Tolerant** **Keep-Alive Logic Control** **Delayed Reset Flag**

Available in TO-263 Surface Mount Packages

Maintain Normal Operation in RF Fields Up to 400MHz at 300V/m Signal Strength

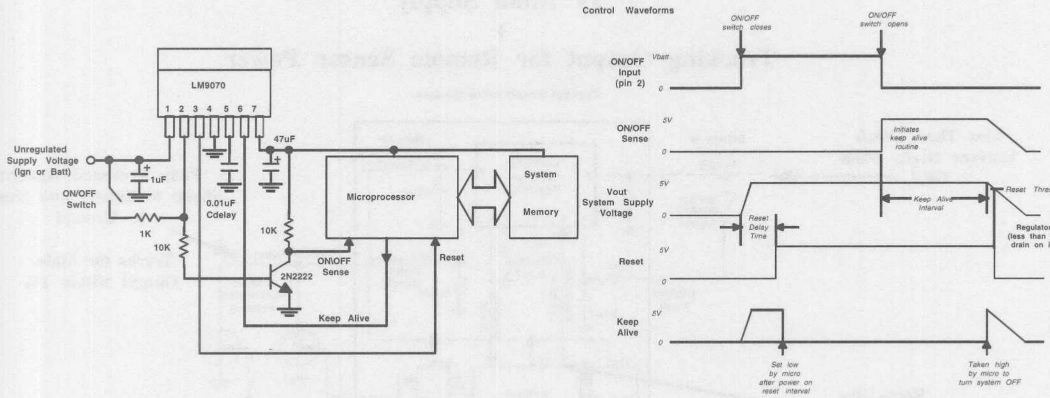
Here is an overview of the three new voltage regulator products. We refer to them as *system* regulators because they contain additional functions beyond the simple three-terminal regulator.

All three contain a delayed reset output flag. This active low output is used to hold a system microcontroller in a reset condition whenever the supply voltage (the regulator output voltage) falls out of regulation for any reason. This ensures proper power supply biasing before allowing the system to execute functions.

While designed primarily for automotive application these regulators contain all of the conventional protection features required. These include reverse battery connection protection, survival during load dump input supply transients, overvoltage protection and a wide operating temperature range. They are also available in the TO-263 surface mount power package and all exhibit immunity to high frequency interference. This ruggedness is also desirable in many non-automotive applications.

The LM9070 and the LM9072 are available in 7-lead and 9-lead packages. The extra pins provide a unique ON/OFF switching control called Keep-Alive control. The LM9071 is a standard 5-lead regulator that provides a programmable delay reset flag output.

Keep-Alive Control Logic



Regulator Remains Biased After System is Switched OFF Until Micro Commands Power-down!

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This illustrates the Keep-Alive logic control. In many systems the microprocessor is required to perform various “house-keeping” tasks when the system is switched OFF. These may include resetting system loads and storing system variables into EPROM which may require an indefinite length of time. Keeping power applied for as long as is required is the benefit of Keep-Alive control.

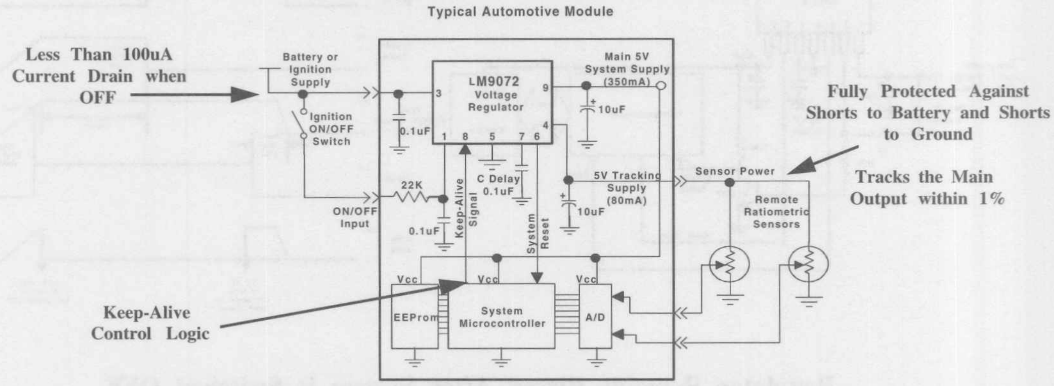
The LM9070 is shown and pin 2 is a conventional ON/OFF input line. This allows the input to be directly connected to the regulator and power ON and OFF controlled by a switch. When OFF the current drain on the input supply is only 20 to 30 µA.

Once switched ON the regulator provides power to the system micro and the micro sets the Keep-Alive control input, pin 6, low. This low level on pin 6 keeps the regulator ON regardless of the state of the ON/OFF input. The NPN transistor shown is used only to signal the micro that the system has been commanded to switch OFF by the opening of the ON/OFF switch.

When switched OFF the regulator remains fully operational. The micro senses the OFF command and can begin whatever is required for a power down routine. This routine can take as long as necessary because the system doesn’t actually turn-OFF until it is ready to do so. At the end of the power-down routine the micro simply takes the Keep-Alive input high and the regulator shuts OFF and reverts to the low current drain standby condition until switched back ON.

The LM9072

5V Main Supply + Tracking Output for Remote Sensor Power



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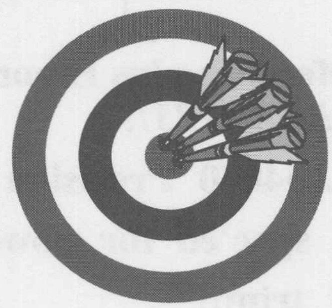
The LM9072 is a dual output voltage regulator which also has the same Keep-Alive control logic as the LM9070.

In addition to the Main 5V, 350mA regulated output there is a second tracking output which can be used for powering sensors and circuits physically located outside the main module. This output voltage is derived from the Main output and tracks it to within a 1% tolerance. This tracking characteristic is important if the supply potential of the remote element must closely match that of the Main system as is the case when powering ratiometric sensors such as potentiometers as shown.

The tracking output is fully protected against not only short circuits to ground but also shorts to the input potential. This is a fault protection feature that is necessary when supply wires travel outside the main module to remote elements where *anything* can happen.

Precision Voltage Control with LM3411

- Better tolerance of Linear, Switching regs
- Control super-LDO at high current
- Drive isolated opto feedback



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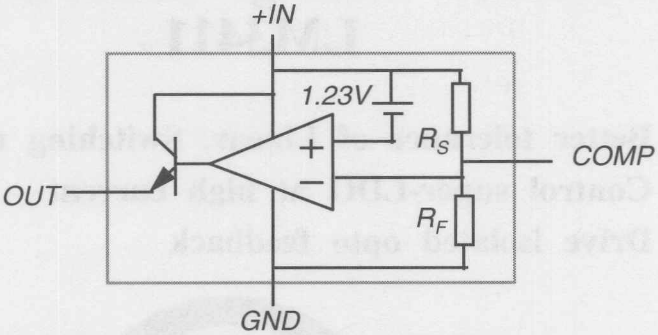
The original application for the LM3411 was in isolated switching power supplies (either DC/DC or offline). Many of these supplies use a PWM controller on the primary side to drive the power stage, which is transformer-coupled and inherently isolated. To provide isolated feedback, a device is used to sense the output voltage and pass a feedback current through an optocoupler, to be received and "decoded" on the primary side. Several other devices have been used to develop this secondary-side feedback, but none have been specified for this function until the LM3411.

Even for non-isolated linear and switching power supplies, the LM3411 can be used to provide tight control of the output voltage. If the power supply has multiple outputs, with only one output directly regulated, the use of the LM3411 will enhance the precision of the main output, and allow the other outputs to have an acceptable tolerance (usually better than +/- 5%) without post-regulation.

LM3411: Flexible Building Block



- **Precision Reference plus Error Amp**
  -- now in SOT23!!
- **Based on LM4040 Precision SOT Reference**
- **Designed & spec'ed for power supply uses**
- **No external trim!!**

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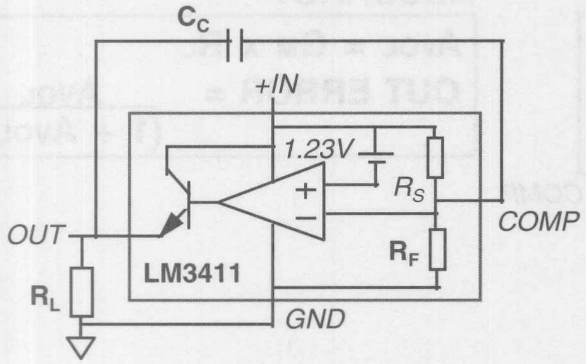
The simplest view of the LM3411 is as a building block, consisting of a precision reference, error amplifier, and an output driver.

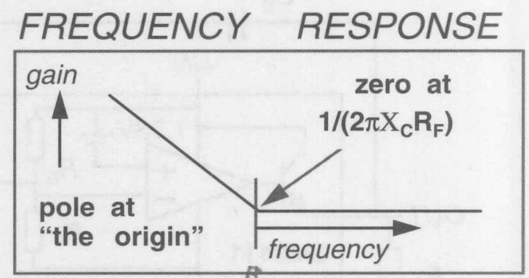
In fact, the LM3411 is a version of the LM4040 precision SOT-23 reference --- modified and specified for use in power supplies. It is specifically intended to provided precision output voltage control, without the use of external or additional trim resistors.

Initially, the device was made available as an 8-pin plastic DIP. However, it is now also available in the 5-lead SOT23-5 package (similar to the SOT-23) for more compact applications.

AC Basics:

- **LM3411 operates as an integrator**
- **C_C & R_F set "zero"**
  - ***make less than reg. bandwidth***





not very accurate point

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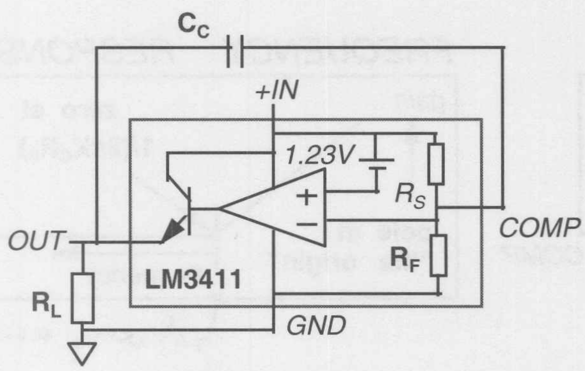
The LM3411 is normally used to generate feedback inside the control loop of a complete regulator or power supply. The frequency response added by the LM3411 is that of a low-frequency integrator (for high DC precision), with a "zero" to shift it to a simple voltage-follower.

An external compensation capacitor C_C around the LM3411 causes the integrating function; C_C working with an internal resistor R_F to produce the "zero" so that, as the regulator frequency response reaches unity-gain, the LM3411 doesn't affect the system stability.

R_F varies from 70kΩ to 118kΩ. Be sure to take this variation into account when selecting the "zero" point frequency. This frequency should be well below your cross over (0dB point) frequency.

DC Basics:

- **G_M x R_L determines accuracy**
  - *sets LM3411 A_{VOL}*



ACCURACY

AVOL = GM x RL

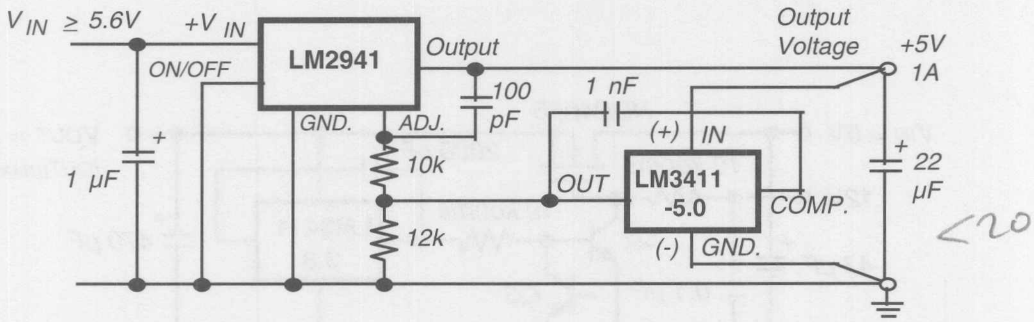
OUT ERROR = AVOL / (1 + AVOL)

The transconductance, or G_M, of the LM3411 affects its control accuracy. As the LM3411 is actually a closed-loop system by itself, G_M x R_L is the open-loop gain of the LM3411 (where R_L is the load resistance of the LM3411). For the 5V part, G_M is typically 3.3mA/mV. If R_L is 300Ω, the LM3411 open-loop gain is about 1000, which is then the maximum low-frequency gain of the LM3411 when used as an integrator. This will contribute an output voltage error of 1 part in 1000 (i.e. 0.1%). If R_L is larger, then this error will be less.

The maximum value of R_L is limited by the LM3411 saturation voltage (how close the output can get to the +IN pin, typically 1V) and the output current range desired. For example, if the maximum output current desired from the 5V part is 1 mA, the maximum value of R_L is:

$$R_{L(MAX)} = (5V\text{-}1V)/1mA = 4\ k\Omega$$

LM2941 + LM3411: 1% tolerance



- **Low-dropout characteristic of LM2941 is preserved**
- **Load regulation improved with remote sensing**

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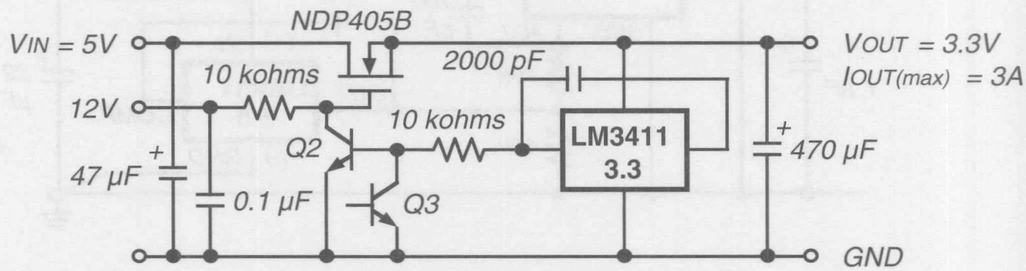
The complete precision 5V, 1A low-dropout regulator is shown above. With the LM2941 providing the power, and the LM3411 providing the precision, the final solution preserves the low-dropout characteristic of the LM2941, but with the 1% accuracy (A grade, over all conditions) of the LM3411. For even better precision with a remote load, the LM3411 output voltage sensing connections can be located at the load, as there is little current in this sensing path to degrade the signal applied to the LM3411.

LM3411 as LDO Controller

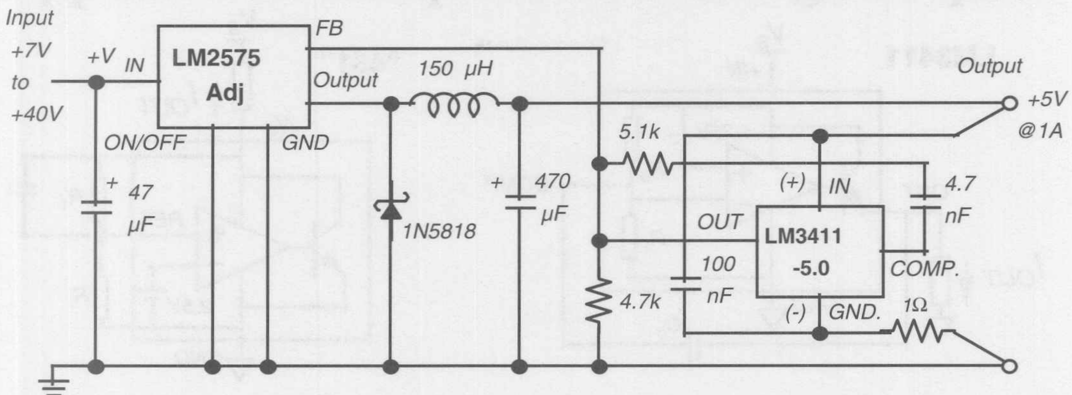


Not only can the LM3411 be used to improve the precision of available linear regulators, it can also be used as the primary control for a linear regulator, even a low-dropout regulator. Having the reference and error amplifier, all it lacks is a power stage. A discrete power stage is all that is needed to complete the regulator.

As this regulator was to be used in a desktop personal computer application, there were two special considerations in the design. The first of these was the need for relative low cost. Although a design of this type usually uses a P-channel MOSFET (because of the available drive), an N-channel device would be less expensive for the same ratings. Fortunately, a 12V supply was available to this regulator, which provided enough gate drive so that an N-channel FET could be used. The device used was an NDP603AL, with an $R_{DS(ON)}$ of 0.16mΩ.

The other special need was for the regulator to be able to directly pass the 5V source to the ouput, in case the load required a 5V supply instead of 3.3V. This was achieved through the use of Q3, which (when pulled high) allows full gate drive to be applied to Q1, putting it into saturation. The load then sees the 5V source through the $R_{DS(ON)}$ of the FET. At full 3A load, the $R_{DS(ON)}$ represents a loss of less than 0.5V.

LM2575 + LM3411: Precision Switcher



- **Use ADJ version of regulator**
- **Add 0.1μF bypass cap, 1Ω resistor to minimize LM3411 noise pickup**

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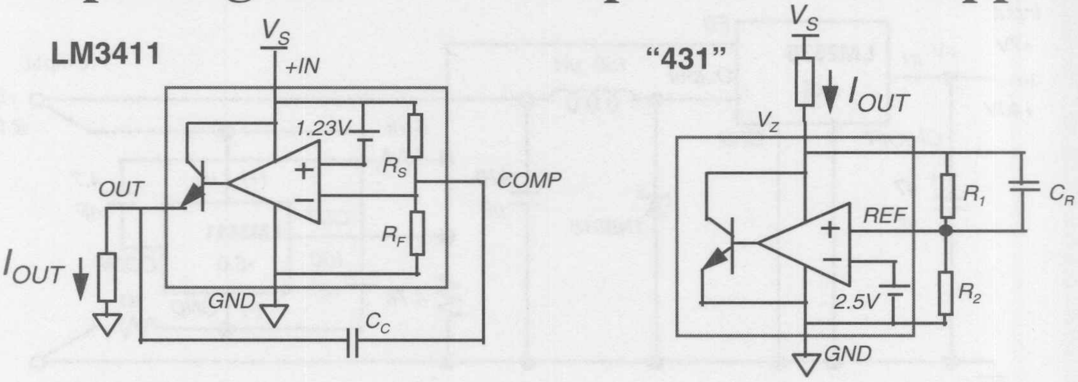
By itself, the LM2575-5.0 is guaranteed to have an output voltage within the tolerance of +/- 5% (over all conditions). However, as in the case of the LM2941, this may not be sufficiently tight when degraded by I x R drops on the way to a remote load, or when this voltage must be used as part of a system reference.

To improve the precision of this switching regulator, the LM2575-ADJ can be used with the LM3411-5.0. (We need the ADJ version of this regulator, as we did with the LM2941, because we need to have separate access to the output voltage and to the feedback node of the regulator.)

At higher load currents, the 100nF-1Ω filter improves the system load regulation, by filtering the high-frequency switching noise from the voltage applied to the LM3411.

Replacing the 431 in opto-isolated supplies



- **431 is normally part of system loop compensation**
- **LM3411 can be set for same frequency response:**

$$C_C \times R_F = C_R \times R_1$$

- **LM3411 output current separate from supply current**

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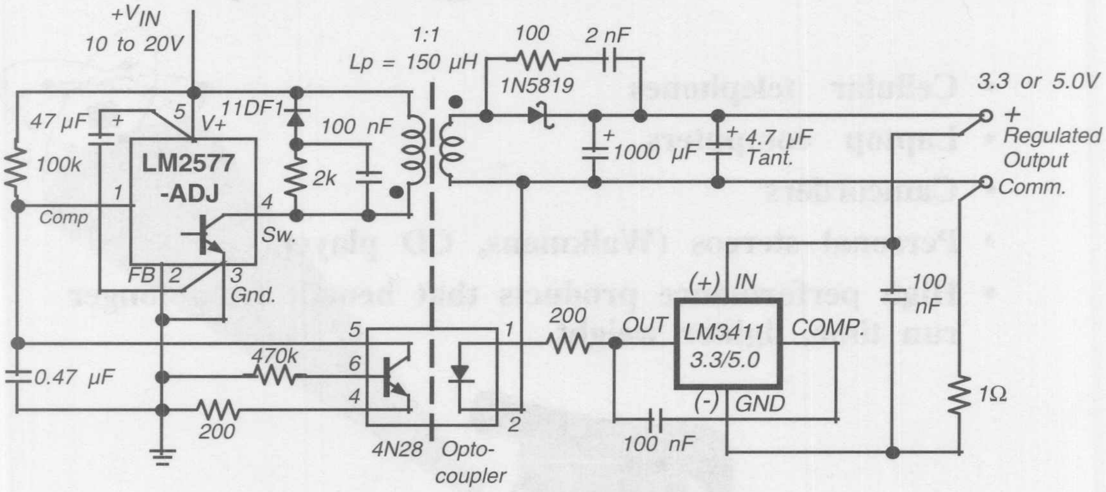
In switching regulator applications, the product most similar to the LM3411 is the older, industry standard 431 (originally from T.I.).

If a power supply has already been designed using the 431, the system compensation will normally have include some compensation using the 431. If we want to replace the 431 with the LM3411, for better accuracy, how do we adjust the compensation?

Fortunately, even though the schematic for the two devices looks a bit different, the net effect is the same - including the effective compensation. The actual compensation capacitor may need to be changed, depending on the resistor values used in the original 431 design.

Another significant difference between the 431 and the LM3411 is the way in which each device is used to drive an optocoupler (as part of an isolated feedback network). The 431 normally has the optocoupler diode on its "cathode" side, where it detects the control current as well as the supply current of the 431. The LM3411 has a separate output pin which is used to drive the optocoupler diode, so that the diode only conducts the control current (which is a function of the output voltage error), not the supply current (which varies with temperature).

Example: LM2577 + LM3411 + Opto

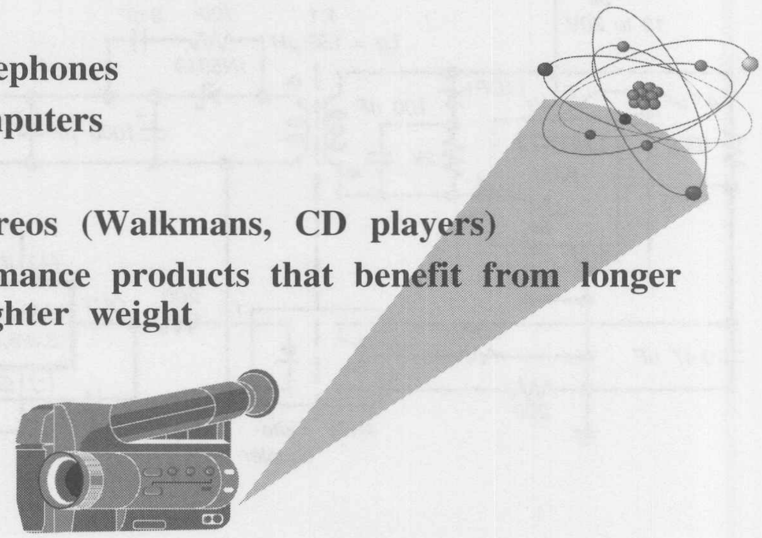


The power supply in the example above was originally designed using the LM385 adjustable reference to provide feedback through the optocoupler to the LM2577 (see AN-777). The LM385 was chosen at that time to show a more precise alternative to the more traditional 431. The application has now been updated further with the use of the LM3411.

With the tight precision of the main output, additional outputs can be developed (with additional windings on the transformer) having tolerance better than 5%.

Li-Ion Cells for High-Density Power

- **Cellular telephones**
- **Laptop computers**
- **Camcorders**
- **Personal stereos (Walkmans, CD players)**
- **High performance products that benefit from longer run time, lighter weight**



The first product available with rechargeable Li-Ion batteries was a Sony camcorder, which boasted the smallest size and lightest weight of comparable models. Other products, such as portable stereo disc players, are also reaping the benefits of smaller size and reduced weight.

The cellular phone market (where ads boast phone weights down to the last tenth of an ounce) is presently incorporating Li batteries into their high end models.

The laptop computer industry is also being drawn to Li batteries by the promise of longer run times. The power consumption of high performance (color display) laptops has so outstripped battery development that there are laptops sold with operating times (between charges) of less than one hour.

Makers of these types of high performance products are becoming very interested in this battery technology that promises more available power.

Energy Densities

Cell Type	Ni-MH	Ni-Cd	Li-Ion
Gravimetric Density (W-Hr/Kg)	55	50	**90**
Volumetric Density (W-Hr/L)	180	140	**210**

- **Gravimetric density (energy per unit weight) is highest for Li-Ion by almost 2 to 1.**
- **Volumetric density (energy per unit size) is closer, but Li-Ion is still the best**

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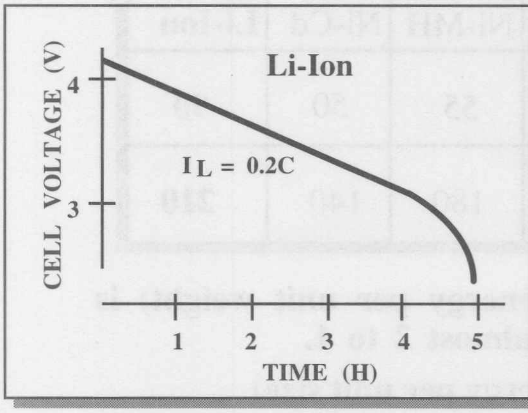
The main advantage using Li-Ion batteries in consumer products is their higher gravimetric density (the amount of stored energy compared to weight). This advantage allows the designer the options of reducing product weight and/or increasing operating time.

As the processors in portable PC's get faster (which means more power hungry) the lure of higher energy will force the use of Lithium.

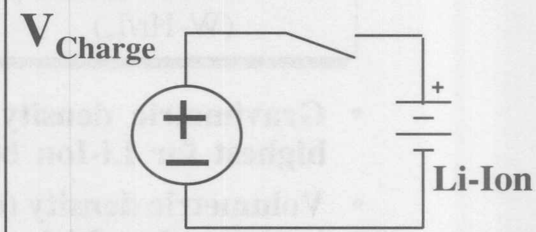
Li-Ion also has an advantage in volumetric density, which means that a reduction in size will also be achieved for a battery of a given Watt-hour rating.

Discharge Profile and Charge Method



Constant Voltage (CV)

- Used for Li-Ion
- End-of-charge detection not required



Li-Ion Cell Voltage Higher, Changes More

30

The higher nominal voltage (about 3.6V) of Lithium is a big advantage over Ni-Cd and Ni-MH for the simple reason that you don't need to stack as many cells in series to get useful voltages.

One Li characteristic that is not as good as the Ni products is the highly sloping discharge voltage. The typical Li-Ion cell will vary from about 4.2V (max) to about 2.2V (min) over the discharge cycle. A varying discharge voltage is an *advantage* in fuel gauging. An accurate measure of battery state-of-charge can be obtained through a simple voltage measurement. However, the non-constant Li voltage requires the design of power converters that operate efficiently over a wide input voltage range - by definition, switching converters are the best choice.

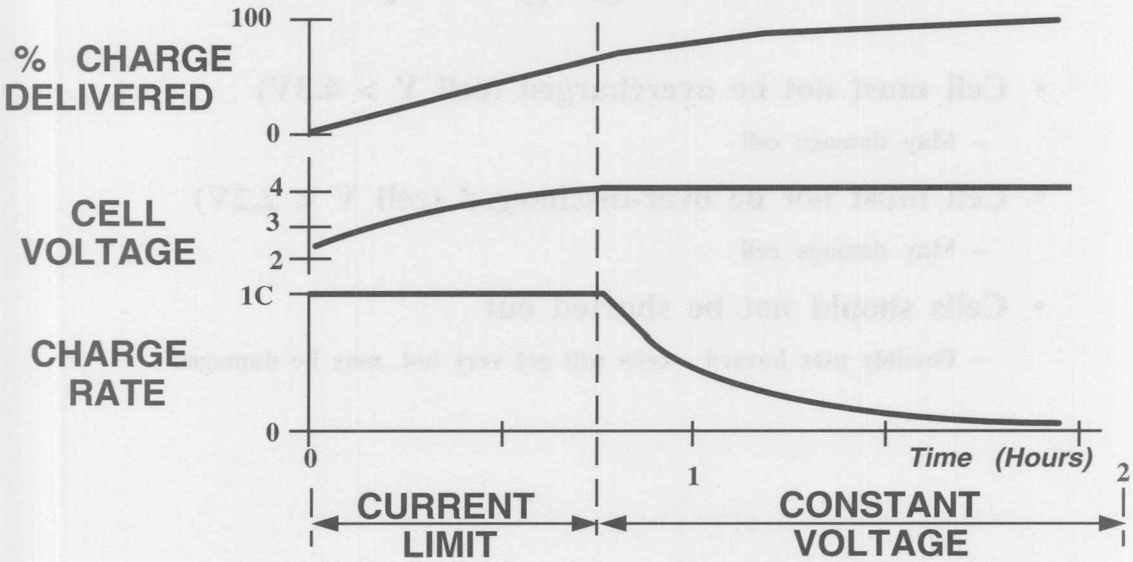
As a result, some manufacturers will be forced to learn a new technology (switching regulator design) that they did not previously use. Although this may be a possible annoyance in the short term, it will eventually yield products with more efficient power supplies.

Charging characteristics of batteries are becoming more important as users demand faster recharging. The minimum standard of acceptance for a good consumer product is typically a one hour fast charge (two hours may be OK).

Fast charging of Ni-Cd and Ni-MH require some rather comprehensive (and expensive) charge termination circuitry to avoid overcharge damage. Li chargers will be a little simpler (and cheaper).

Two completely different charging methods are used for Ni and Li batteries. A constant-current technique is best for Ni, but a constant-voltage charger must be used for Li. The different charging characteristics of the chemistries is a problem, as most equipment manufacturers would like to do a seamless transition from Ni-MH to Li, which would require charging circuits that accept both battery types.

C-V Charge Profile



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A typical C-V(constant voltage) charge cycle is shown above with charging current limited to 1c (which is a charge rate that is equal to the A-hr rating of the battery). During the current limit phase of the charge cycle, about 65% of the total charge is delivered to the battery at the c rate. It is necessary that the charger limit the maximum current during this phase to a value that is safe for the battery.

The constant voltage phase begins when the cell reaches the set voltage (4.2V in this case) and the charger starts reducing the current to hold the battery at this voltage. The constantly-decreasing current during the constant voltage phase means that it takes about twice as much time to deliver the final 35% of charge, compared to the time it took to put in the first 65% during the current limit phase.

This is the source of the one negative side of the C-V charge method: For a given maximum charging current, C-V charging takes about twice as long as C-C charging.

Li-Ion Charging Requirements

- **Cell must not be overcharged (cell V > 4.3V)**
  - **May damage cell**
- **Cell must not be over-discharged (cell V < 2.2V)**
  - **May damage cell**
- **Cells should not be shorted out**
  - **Possible user hazard - cells will get very hot, may be damaged**

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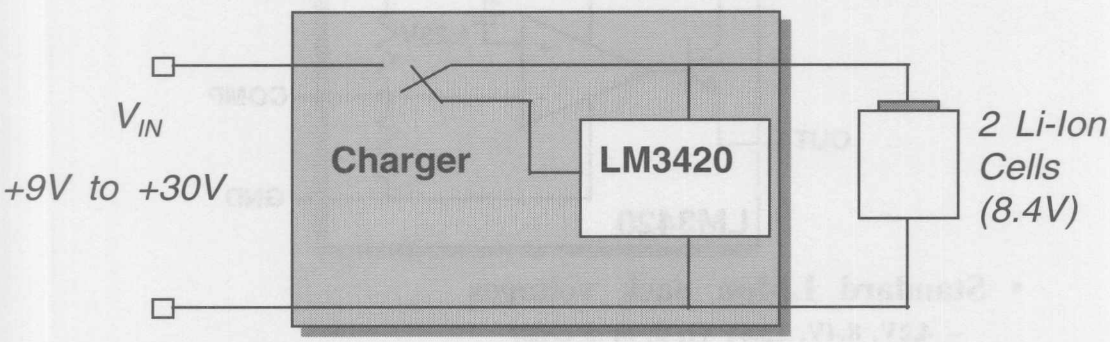
Li-Ion cells can easily sustain damage through "typical" battery mis-use: overcharge and over-discharge. To prevent damage when charging, the cell voltage should never be allowed to rise above 4.3V. At the other extreme, the cells should not be allowed to discharge below a cell voltage of about 2.2V. Operation beyond these limits will cause a non-reversible degradation within the cell that reduces its performance (or kills it altogether).

Another potential problem could arise if the cells are shorted out: they carry enough energy that the cells can get extremely hot.... hot enough to burn a user.

Because of these characteristics, the battery maker shipping Li-Ion cells in consumer products uses an "isolation" technique for product reliability and user safety: The batteries are sold in packs (not single cells) and the packs contain protection circuitry that isolates the cell connections from the battery pack terminals to prevent these problems from damaging the battery or the user.

In addition, the charger for a Li-Ion cell pack must incorporate safety features to prevent over-voltage or over-discharge of the cells.

Lithium-Ion Charge Controller

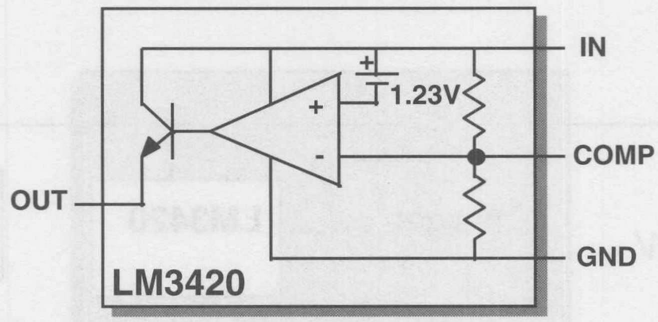


33

A constant-voltage charger for Li-Ion cells includes a device, such as the LM3420, which monitors the cell voltage. This device also controls a pass device to limit the charge current near end-of-charge.

A power source is also key, as it must provide enough energy to charge the cells at the fastest-desired rate. The power source must incorporate over-current protection, to limit the charge rate to a safe level.

LM3420 Matches Li-Ion Charging Needs



- **Standard Li-Ion pack voltages**
  - **4.2V, 8.4V, 12.6V (1, 2, or 3 cells)**
- **Precision charge control meets Li-Ion requirements**
  - **A-grade_+0.5%, Standard precision_+1%**
- **Tiny SOT23-5 package for compact designs**

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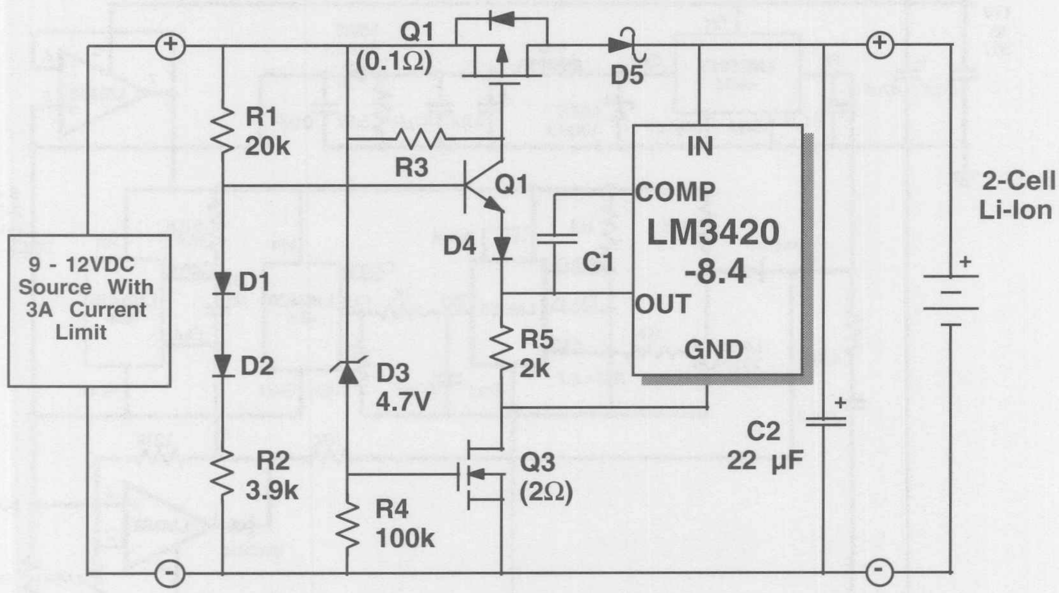
The LM3420 incorporates the essential needs of Li-Ion charging into a single IC:

The main challenge of Li-Ion charging is the tight accuracy requirement for the set point voltage. In a typical discrete design, this requires the use of a voltage trim to get the voltage within the target range. , With a voltage accuracy of 0.5% (room temperature) and 1% over the full temperature range for "A" grade units, the LM3420 eliminates the need for external trims

The LM3420 operates as a “shunt” regulator, which means that the output will begin sourcing current when the voltage applied between input and ground reaches the set voltage of the part.

Minimum board space is used by the SOT23-5 package, and voltage options of 4.2, 8.4, and 12.6V are provided to accommodate 1, 2, and 3 cell charger designs.

3 Amp Li-Ion Battery Charger



Q1	MMBT3904	D1/D2	BAV99
Q2	NDT454P	D3	MMBZ523DB
Q3	NDS7002A	D4	MMBD914

The charger for a two-cell pack requires a full-charge voltage of 8.40V, with an overall accuracy better than 1%. The maximum charging current is 3A (and must be limited by the DC input source). In the typical (constant-voltage) mode of operation, the LM3420 is a controller in a feedback loop that precisely regulates the voltage of the batteries to 8.4V. With the built-in precision of the LM3420 control voltage, external voltage setting resistors are not needed.

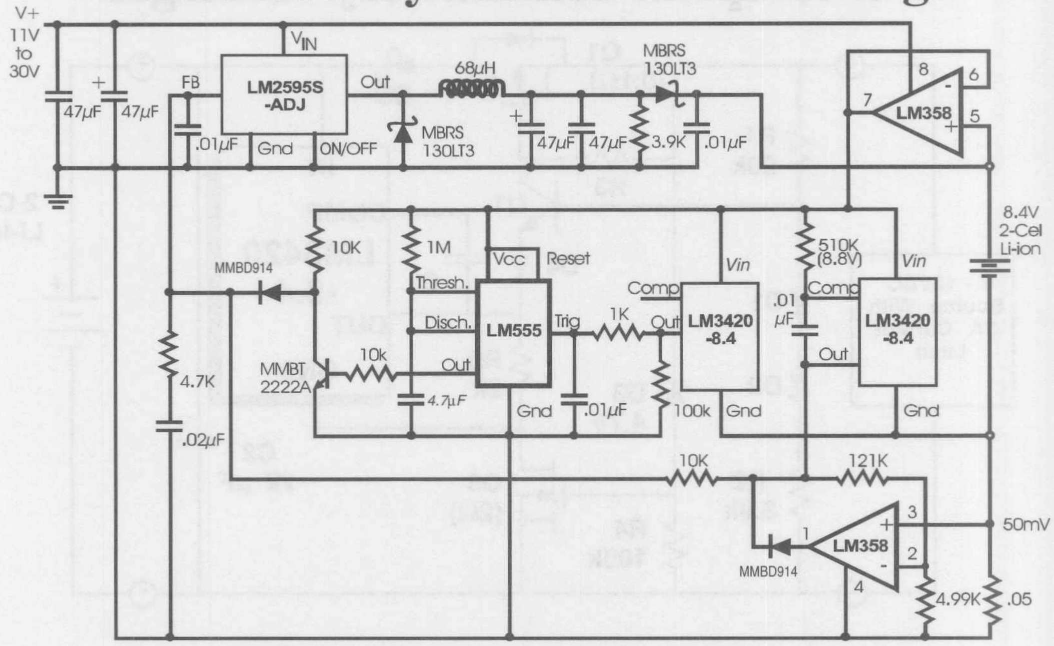
Charge current is adjusted by control of Q2, a P-channel FET. The drive to Q2 is the voltage across R3, which conducts the difference of a 300 µA bias current (developed by R1, D1, D2, Q1, and R5) and the output current of the LM3420. When the battery voltage is below the target of 8.4V, the LM3420 output current is near-zero; this applies maximum drive to Q2. In this mode, the charger delivers the full current available from the source to the cells. The DC source must provide current limit at about 3A. To minimize self-heating, Q2 should be chosen for low on-resistance.

When the battery voltage reaches 8.4V, the LM3420 regulates the battery voltage by sourcing current, which adjusts the Q2 gate voltage as required to hold the battery voltage at 8.4V. In this constant-voltage mode of operation, Q2 operates in its linear region in response to the feedback from the output of the charger circuit (through the LM3420).

If the batteries are not being charged, or the charger does not have enough voltage applied to work correctly, the charger must disconnect itself from the cells to prevent energy drain. As long as a DC source voltage greater than 4.7V is applied, an "on/off switch" made up of Q3, R4, and the 4.7V zener allows the circuit to operate. If the DC input source is removed, Q3 will turn off and reduce the drain on the batteries to less than 1 µA (which is much less than the self-discharge rate of the cells).

The Schottky diode, D5, prevents battery drain due to current flowing back through the internal diode of Q2 when the DC input is removed. A 16A diode was selected to minimize power losses, but a 6A could be used for cost savings.

Li-Ion Battery Pulsed Fast Charger



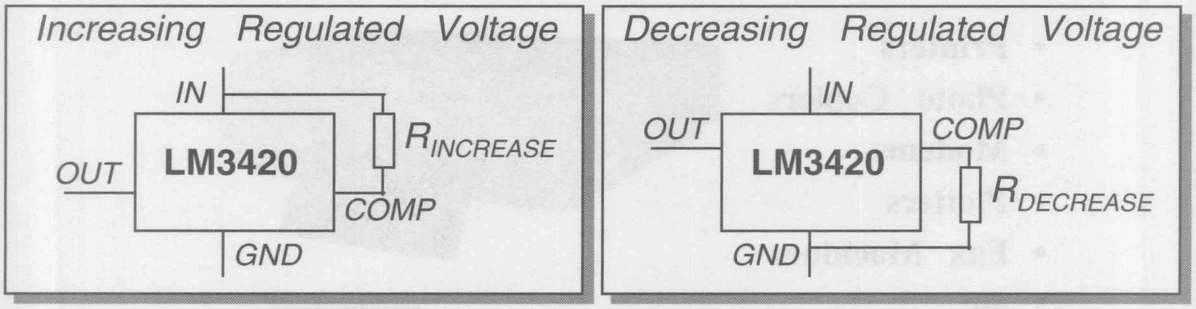
This circuit reduces charge time from 2 hours to 1.25 hours. It accomplishes this charge rate by temporarily increasing the voltage applied across the battery to overcome internal impedance and achieve a higher charge current.

When the circuit begins charging a completely dead battery, the circuit runs in a constant current mode. One of the LM358 op-amps compares the voltage across the 0.05 ohm resistor and limits the current to 1 amp through the feedback pin on the LM2595 switching regulator. As the battery charges the voltage across it will increase. When the voltage reaches 8.4 volts, the LM3420-8.4 (the left hand one) will start the LM555 timing circuit. When the timing circuit triggers (delay time based on the 4.7uF and 1Mohm resistor), the input to the MMBT2222A will go low and turn off the transistor. This causes the input to the feedback pin on the LM2595 to go high and turn off the switching regulator. With the supply removed from the battery, it's voltage will begin to fall. As it falls below 8.4V, the transistor at the output of the LM555 will turn on and the circuit will begin charging again. This off/on operation will occur very quickly early in the charge process, but as the battery becomes more fully charged, the off time will increase.

The LM3420-8.4 on the right lmits the maximum voltage across the battery to 8.8 volts. This higher voltage is set by the 510K resistor across the COMP and Vin pins of the LM3420-8.4.

Some other features include the LM358 in the upper right which acts as a buffer between the battery and the controll circuitry. This prevents discharging the battery when the charge circuit is idle. The MBRS130L in series with the switcher ouput performs the same function.

Adjusting the Regulated Voltage



- **Adjust V_{REG} up to ±10% with external resistor**
- **Equations given in datasheet**

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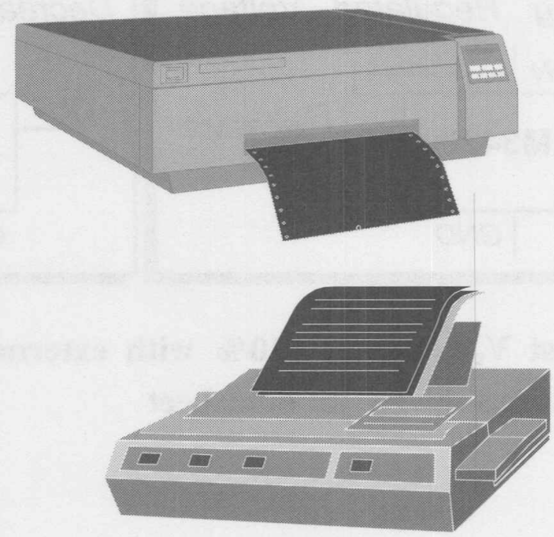
While the standard voltage for a Lithium-Ion cell is 4.2V (fully charged), other batteries are being developed that have a slightly different voltage. The set point of a charger based on the LM3420 can be adjusted LM3420 using a single resistor. The adjustment range is +/-10%.

Power Conversion In Line-Powered Equipment

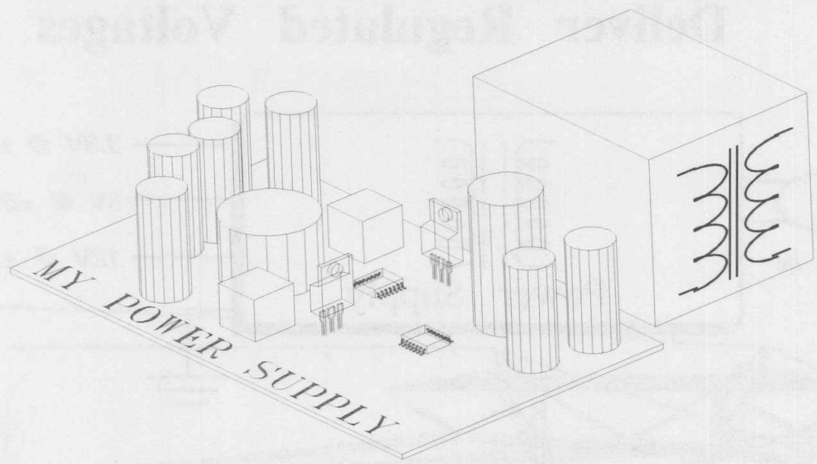
- **Printers**
- **Photo Copiers**
- **Modems**
- **Plotters**
- **Fax Machines**
- **etc.**



Most equipment used in the office or home draws its power from the AC line. This line may be between 90 - 264VAC and 47 - 63Hz, depending on which portion of the world the equipment is located. However, the internal circuitry and motors rarely operate at these voltages. It now becomes necessary to provide Power Conversion within the equipment. This conversion means a power supply.

The following pages will show some examples of the power needs within line powered equipment and methods to solve those power needs.

You Can Design Power Supplies !



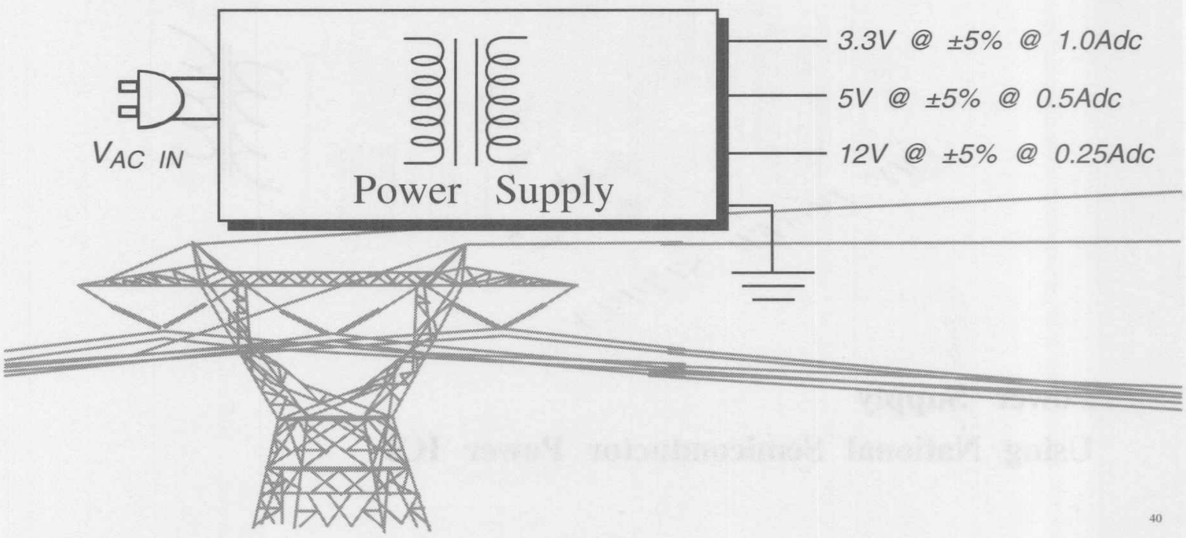
Power Supply
Using National Semiconductor Power ICs

The power conversion technique we will address includes a line transformer which provides safety isolation from the line voltage. The line transformer is the only portion of the power supply which needs to meet EN60950 (IEC950) standards. All voltages produced by the transformer will be assumed to meet the requirements for SELV (Safe Extra Low Voltage) circuits. These transformers may be purchased as catalog items, or may be specified with custom requirements.

The power supply following the isolation transformer can be easily designed using National Semiconductor's Power ICs. Standard data sheets include design procedures and application hints on the use of all parts. Other parts will be supported with design software providing a complete solution.

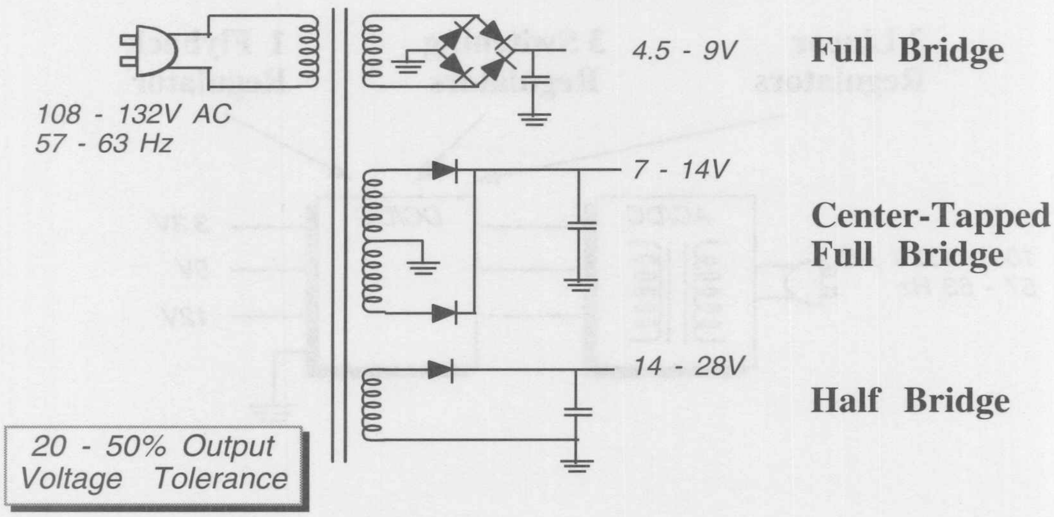
Now you can confidently design the power supply for your own equipment.

Power Supply Task: Deliver Regulated Voltages



Above is an example of a power supply requirement for any given piece of line powered equipment. The AC input line must be converted by the power supply to provide three DC output voltages which are well regulated over input line and output load. Most circuitry requires low output ripple on its supply voltage. Output ripple is generally specified to be no more than 1% of the output voltage.

AC Line Rectification



One must understand the characteristics of low frequency line transformers to begin to design a power supply.

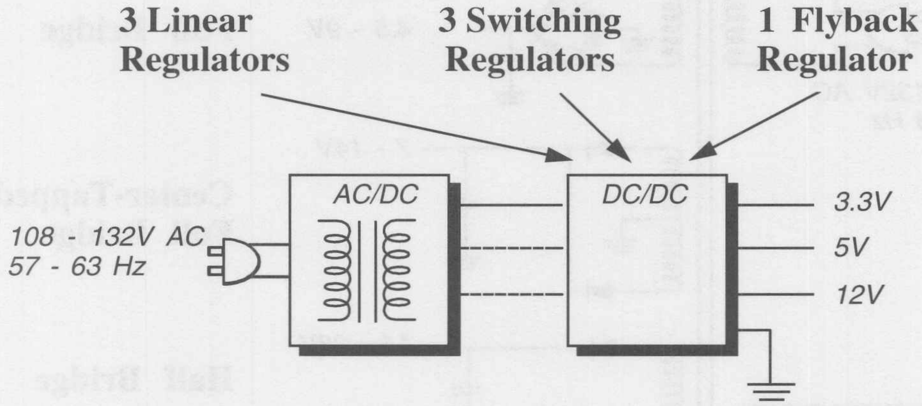
The figure above shows three different rectification techniques to develop a basic DC voltage source. The first, Full Bridge, is most efficient in the transformer, but requires a four diode rectification bridge. The second, Center-Tapped Full Bridge, requires more turns of wire in the secondary, but saves two diodes. The output ripple on both Full Bridge designs is the same. The last, Half Bridge, is a low cost method best for low output power. The half bridge will result in poor utilization of the transformer and high output ripple.

Output regulation is poor in an line transformer. If one assumes an ideal transformer, the regulation of the output will be no better than the input. Typical design requirements for input voltage is nominal ±10% (some require ±12.5%) . In the case above, nominal is 120VAC.

Therefore, the output voltage tolerance is no better than ±10%. Now add resistance of the transformer windings. This results in output voltage variation with load. The change in output voltage with load is a complex function and not intended for this discussion. In general, load regulation can vary from 10 to 40%, with 25% being typical load regulation for a reasonable size transformer.

In general, the size of the transformer is a function of a number of items: output power, load regulation, efficiency, maximum allowable temperature rise, and cost (core material is the primary variable).

DC-DC Converter Options



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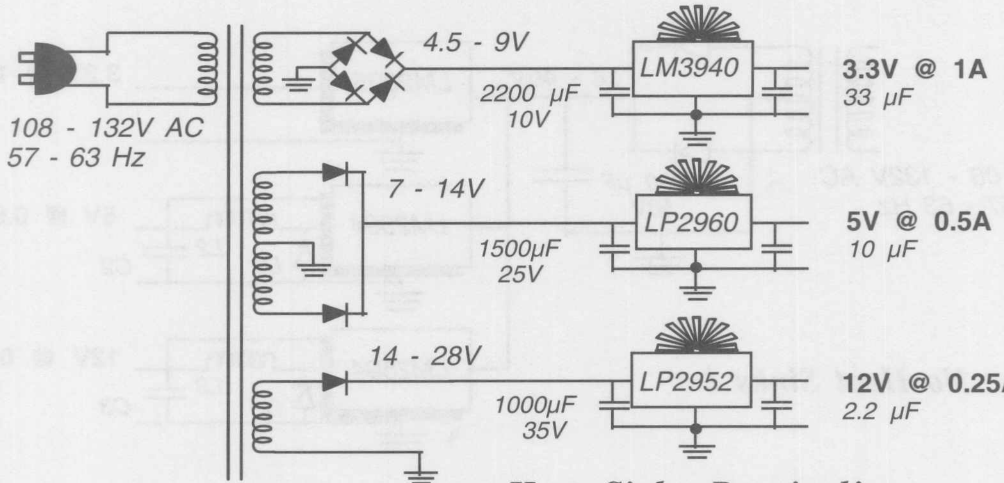
National Semiconductor has three ways to provide the tightly regulated DC outputs required in this example.

The first is using three linear regulators from three different transformer outputs. This is the lowest efficiency solution. However, it is also the easiest to design and has the lowest component count. This solution requires an output winding for each supply to obtain acceptable power loss in the linear regulators.

The second is using three switching regulators (buck DC/DC) from a single transformer output. This provides excellent efficiency. However, the design complexity increases slightly and component count increases. This solution requires only a single poorly regulated output from the transformer.

The third is a single flyback switching regulator. This also provides good efficiency, but is even more complex. The example above has no compelling reason to use a flyback regulator. A more appropriate set of requirements for this type of switching regulator will be presented later.

Linear Regulator Design



Heat Sinks Required!

The solution above shows how one would use linear regulators to obtain the desired regulated output voltages. The bulk capacitors across the rectified transformer windings may be reduced depending on the transformer being used. The linear regulators provide exceptional of the 120Hz ripple voltage found at the output of the transformer. The output capacitors are selected using the Application Hints in the datasheet for each part.

Each linear regulator must have a heat sink to prevent over heating. All calculations for power loss in the linear regulators will be performed at 132VAC input to the transformer and full load on the outputs. The power loss in the linear regulator can be described as:

$$P_{LOSS} = Vin \times I_{GND} + (Vin - Vout) \times I_L$$

Where Vin is the average DC from the transformer, I_{GND} is the ground pin current at full load (see data sheet), and I_L is the output current.

The LM3940 is the low dropout regulator chosen for the 3.3V output at 1Adc. Vin is approximately 5.5Vdc, and the calculated power loss is 2.8W (I_{GND} = 110 mA).

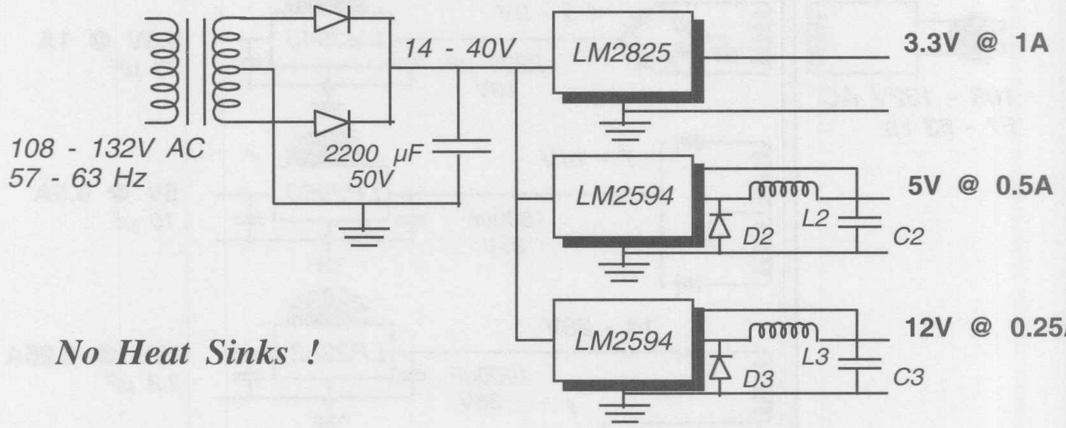
The LP2960 is the low dropout regulator chosen for the 5.0V output at 0.5Adc. Vin is approximately 7.3Vdc, and the calculated power loss is 1.3W (I_{GND} = 21 mA, max).

The LM2952 is the low dropout regulator chosen for the 12V output at 0.25Adc. Vin is approximately 15.9Vdc, and the calculated power loss is 1.3W (I_{GND} = 21 mA, max).

The total output power is 8.8W and the total power loss is 5.4W. As a result, the line transformer must provide 14.2W. If lower cost non-LDO were used, there would be an additional 3.8W loss (because of the higher input voltage required, countered by lower supply current). This would result in a 27% increase in transformer size.

If a lower-cost transformer with a single winding was used, even with LDO regulators, the extra loss would be 14.9W. As a result, 3 windings are required.

Switching Supply Provides Better Efficiency



No Heat Sinks !

Above is shown a solution using three switching regulators. The average voltage supplied by the line transformer is 14 - 40V. This voltage represents a relaxation of the the line transformer regulation and can result in a smaller transformer. Again, the size of the input capacitor may be reduced and 120Hz ripple rejection is excellent.

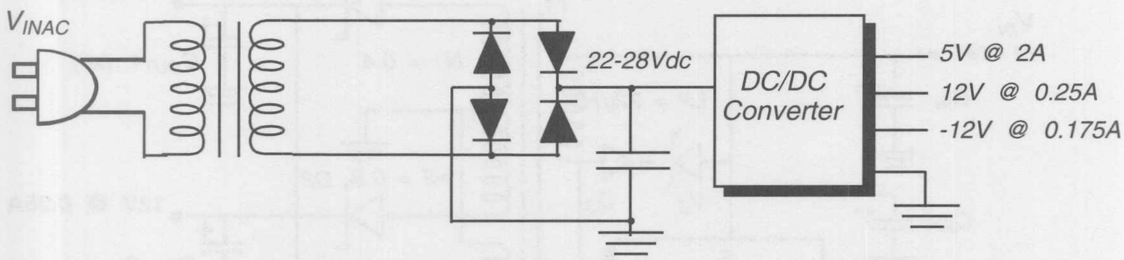
None of the switching regulators requires a heat sink given proper heat sinking to the PCB (Printed Circuit Board).

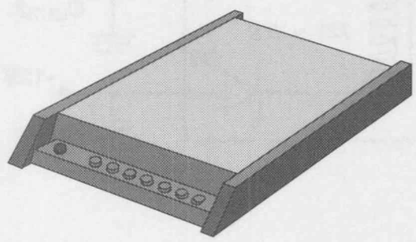
The components are selected by a software design tool called Switchers Made Simple (ver. 4.2) which is discussed later in the presentation.

L2	150uH	
C2	82uF	UPL1V820MAH (Nichicon)
L3	220uH	
C3	82uF	UPL1C820MAH (Nichicon)

The overall efficiency of this switching regulator solution is approximately 80%. Therefore, the line transformer only needs to provide 11W. The line transformer used in this solution can be 22% smaller than that used in the previous linear regulator solution. Also, only one output winding is required, thus simplifying the transformer manufacture.

Modem Supply Requires Multiple Outputs

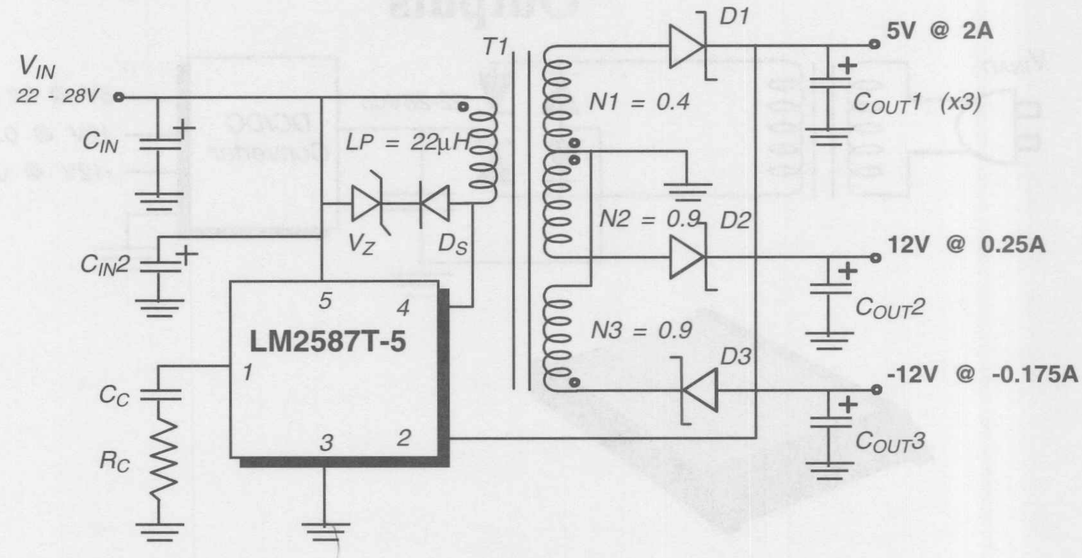




The design above, for a modem application, required a DC/DC converter that was easy to configure and had low power loss. To meet these needs, a multi-output flyback switching regulator was used to regulate the voltage developed by the line transformer. Not only does the use of a switching regulator keep the power loss much lower than if a linear regulator were used, the flyback topology is well suited for providing negative voltages while using one of the positive outputs as the feedback point.

The easiest way to develop such a DC/DC converter is to generate a SIMPLE SWITCHER converter design with the software tool "Switchers Made Simple."

Instant Design with Simple Switcher



Part Summary as generated by Switchers Made Simple 4.2:

U1	5.00 A	National	LM2587T-5
T1	(Software provides detailed specification...)		
C_{IN}	270.00 µF*	Nichicon	UPL1J271MRH
$C_{IN}2$	100.00 nF	AVX	SR595C104KAA
$C_{OUT}1$	2.70 mF (x3)	Nichicon	UPL1V272MRH
$C_{OUT}2$	330.00 µF	Nichicon	UPL1V331MPH
$C_{OUT}3$	270.00 µF	Nichicon	UPL1V271MPH
R_C	3.00 k Ω	Dale	CCF-07302J
C_C	330.00 nF	AVX	TAPA334K035R
D1	Schottky	Motorola	MBR745
D2	Schottky	Motorola	MBR1100
D3	Schottky	Motorola	MBR1100
VZ	20.00 V	Motorola	SA20A
D_S	Ultrafast	Motorola	MUR120

*May require a larger value if used as the bulk capacitor for the line transformer.

This DC/DC convertor is approximately 75% efficient.

Undervoltage Lockout

- **Special requirements when using line transformer**
  - **Full load vs. no-load voltage!!**
  - **Need to have large hysterisis to avoid "motor boating"**
- **Graceful shutdown during brown-out conditions**
- **Clean power up**

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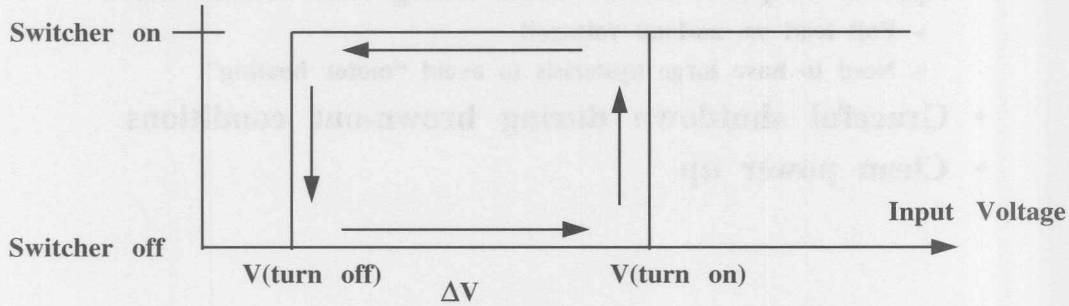
Although the power supply is designed to operate properly over a given input voltage range, there is no guarantee that the line voltage available to the customer will always stay within that range. For over voltage and transient conditions, we can protect our power supply and other internal circuitry by using a zener diode to clamp the input voltage. For brown out conditions we can use an undervoltage lockout circuit in conjunction with the shutdown pins on the Simple Switchers.

The undervoltage lockout must have special features due to the load regulation of the 60Hz transformer. If we were to set a fixed on/off voltage for the power supply the following would occur;

1. The input voltage falls below our minimum operating point (say 105Vac).
2. The undervoltage circuit trips and turns off the power supply and equipment.
3. The 60Hz transformer is unloaded and the output voltage jumps up above the on/off trip point.
4. The power supplies try to start and pulls the output of the 60Hz transformer down again.
5. Steps 2 through 4 repeat constantly causing the power supplies and equipment to oscillate on and off.

The problem is the difference between no-load and loaded output voltage of the 60Hz transformer. Our undervoltage lockout must be able to turn on the power supplies once the input voltage is within our operating range (about 108Vac). But, once the supplies turn on, recognize that the output voltage will be pulled down, and not turn off until the input falls below our operating minimum (about 105Vac).

Undervoltage Lockout - Hysteresis



- **V(turn on) = No load voltage from transfomer @108Vac**
- **V(turn off) = Full load voltage from transformer @105Vac** 100
- **ΔV = V(turn on) - V(turn off) ; Hysteresis**

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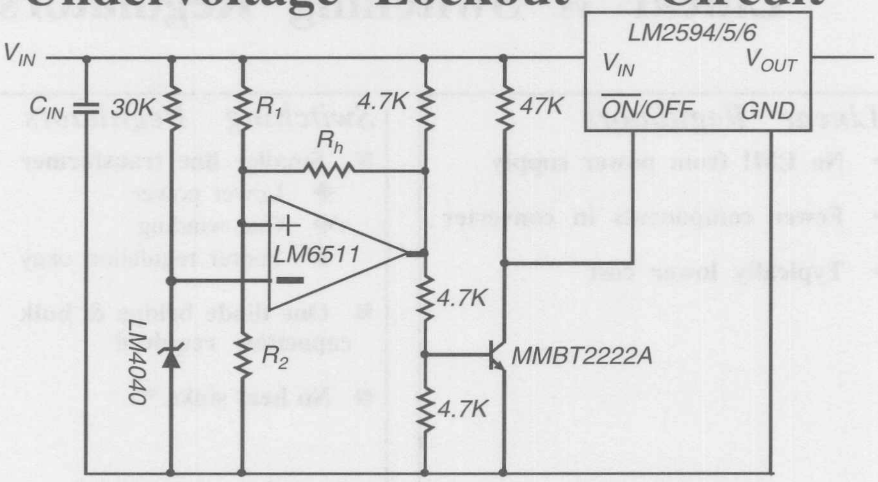
By using an undervotage circuit with hysteresis, we can prevent the power supply from oscillating on and off.

V(turn on) is the no load voltage from the transformer. This is very close to $108 \cdot \frac{Ns}{Np} \cdot \sqrt{2}$

V(turn off) is the full load voltage from the transformer at 105Vac. This value is very sensitive to the load conditions for each design, and should be measured on the bench during design.

ΔV is the difference between these two voltages, and is the value we will use for the undervoltage lockout hysteresis.

Undervoltage Lockout - Circuit



$$V(\text{turnon}) = \text{Vref} \bullet \frac{R_1 + R_2}{R_2} \qquad \Delta V = \frac{1}{1 + \frac{R_h(R_1 + R_2)}{R_1 \cdot R_2}}$$

The schematic and equations above will provide a hysteretic undervoltage lockout circuit design. Some assumptions have been made to simplify the equations, primarily that $R_h >> R_1$ and R_2 . Also, the current through R_1 and R_2 should be at least 10 times greater than the input current to the comparator. The LM6511 input current is typically less than 50nA (datasheet maximum over all conditions is 200nA which occurs at -40°C).

Once the circuit has been incorporated with the complete power supply and powered circuitry, the values of R_h, R_1, and R_2 can be optimized.

Linear vs. Switching Regulators

Linear Regulators	*Switching Regulators*
• **No EMI from power supply** • **Fewer components in converter** • **Typically lower cost**	■ **Smaller line transformer** ◆ Lower power ◆ One winding ◆ Poorer regulation okay ■ **One diode bridge & bulk capacitor required** ■ **No heat sinks** * * Small heat sinks may be required at higher ambient temperature or output power levels

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Each type of regulator has its own strengths and weaknesses.

The linear regulator provides a fast simple solution. Overall, the linear regulator will be the lowest cost solution. The drawbacks are a complicated line transformer with multiple output windings, low efficiency, and heat sinks to dissipate the power lost in the regulator.

Switching regulators provide a more efficient solution at the expense of complexity. The line transformer can be less expensive than that used for the linear regulator. One benefit of higher efficiency is the ability to omit heat sinks unless the ambient temperature is very high (>50 °C at rated current).

One issue which may arise is EMI from the switching regulators getting back into the AC line. Linear regulators do not generate any EMI. Switching regulators may require a filter stage at the output of the line transformer if the bulk capacitor does not provide sufficient attenuation at the switching frequency.

While new requirements for power factor correction (PFC) are being put in place in Europe, they do not generally apply to the sub-50W power supplies we are discussing here. For this reason, none of the solutions here are intended to address PFC.

Why Use A SIMPLE SWITCHER™ DC/DC Converter?

- **Simple to design-in**
- **Guaranteed system specifications**
- **Few external components**
- **Standard magnetics**
- **Software support**

While there are many types of DC/DC voltage converters on the market, there are few that combine the ease-of-use and adaptability of the SIMPLE SWITCHER DC/DC converters.

SIMPLE SWITCHER products offer guaranteed system specifications, such as maximum output voltage tolerance, not just the tolerance of a subsection of the integrated circuit.

In addition, a SIMPLE SWITCHER is easy to configure, with a variety of standard output voltages available. A few external components are required, and they are fully specified in the product documentation. Components which may be unfamiliar the the system designer, such as magnetics, are available as standard part numbers from other vendors.

Design software is also available, to customize a SIMPLE SWITCHER converter for a specific application.

Power Supply Solutions for Line-Powered Equipment

- **Multiple-output DC/DC Converters**
  - **– SIMPLE SWITCHER Flyback Regulators**
    - **» 10-25W Power: New LM2585, LM2586, LM2587, LM2588**
- **Single-output DC/DC Converters**
  - **– Low-Dropout Linear Regulators**
    - **» 0.05 - 0.5A Loads: LP2980/1, LP295x, LP2960**
    - **» 5 to 3.3V at 1A: LM3940**
  - **– SIMPLE SWITCHER Regulators**
    - **» Step-down, ≤ 3.0A Loads: New LM2594/5/6, and LM2597/8/9**
    - **» Step-up, 10-25W Power: New LM2585/7, LM2586/8**

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We have seen three different ways to provide regulation of a multi-output line-powered supply.

•A single-output line transformer can be used with a multi-output switching regulator, such as the new SIMPLE SWITCHER flyback converters LM2587 and LM2588. This yields a relatively simple, high-efficiency supply.

•A multi-output line transformer, with output voltages close to the desired levels, can be used with low-dropout linear regulators on each output. This yields a low-component-count power supply.

These LDOs may include:

LP2980/2 for < 50 mA loads

LP2950/51 for < 100 mA loads

LP2952/53/54/57 for < 250 mA loads

LP2960 for < 500 mA loads

LM3940 for < 1Amp loads

•A line transformer with either a single output or multiple outputs can be used with single-output switching regulators, such as the new SIMPLE SWITCHER buck regulators LM2594, LM2595, LM2596, LM2597, LM2598, and LM2599 and the existing line of SIMPLE SWITCHER buck regulators LM2574, LM2575, and LM2576. This yields a high-efficiency power supply with independent control of each output.

LM2594, LM2597
LM2595, LM2598
LM2596, LM2599

New SIMPLE SWITCHER™ Buck Regulators

New Buck Regulators Add to SIMPLE SWITCHER Line

	Basic	Extra Features
0.5 Amp	LM2594	LM2597
1 Amp	LM2595	LM2598
3 Amp	LM2596	LM2599

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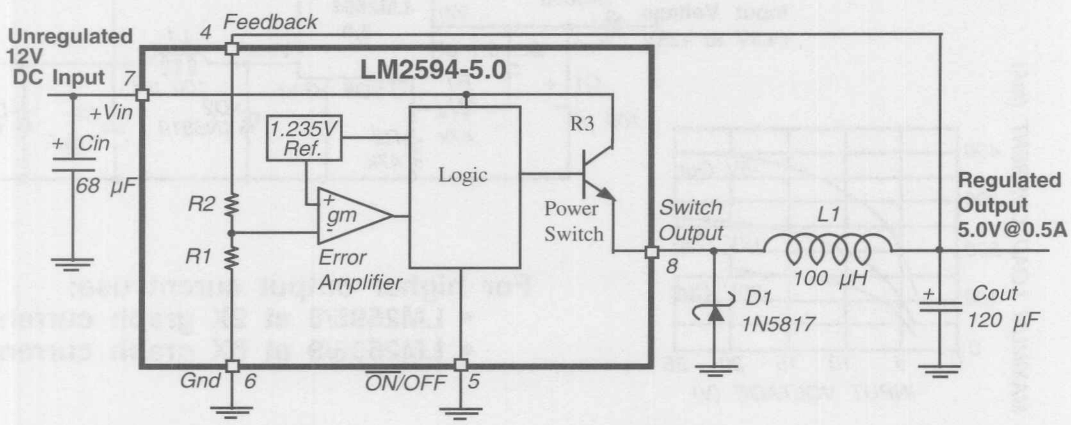
This new series of SIMPLE SWITCHER converters is a full line of step-down switching regulators, with three load current ranges including up to 3A. For each load current range, there is a "basic" and a "full-featured" version. The basic regulators offer a complete power converter solution (when used with the four external components specified in the datasheet or design-in software). Additional functions are available with the full-featured products, which allow further power supply design flexibility.

Note about nomenclature:

In this introductory material, this complete converter family will be referred to as "LM259x." Where special characteristics of the different versions are described, the products are referred to by their individual part numbers.

Basics of a Buck Regulator

- **Step down switching voltage regulator**
- **Vin > Vout (e.g. : Vin = 12V, Vout = 5V)**



A buck or step-down regulator is a switching voltage regulator, which is used in applications where the input voltage is higher than the output voltage (hence the name step-down).

The most important components in a buck regulator are the power switch, the error amplifier, the reference, the resistor divider, and the logic. A negative feedback loop is used to regulate the output.

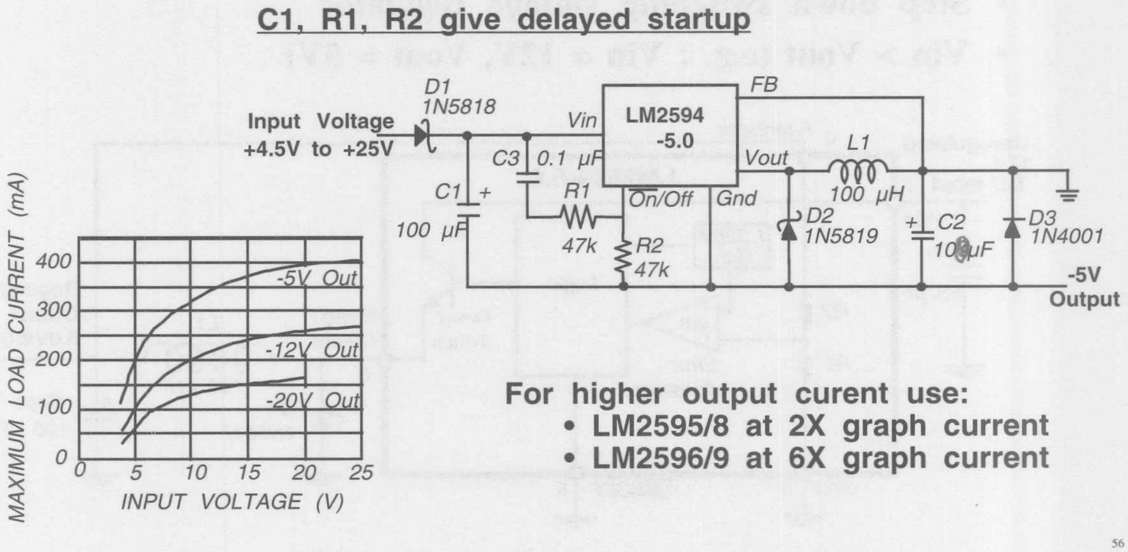
For example, we have a 12V input and a 5V@0.5A output. This means that Vout average is 42% of Vin average. The duty cycle of the switch will be approximately 5/12= 42%, so the switch is ON 42% of the total time.

To see the effect of the feedback control, let's exaggerate a little and assume that the output goes up to 5.5V. The feedback pin will measure this and the voltage at the -input of the error amplifier will be higher than 1.235V (which it is in equilibrium (at 5V), because of the resistor divider), thus higher than the reference voltage. The error amplifier will amplify the difference and send this signal to the logic part of the circuit. The logic drives the base of the power switch, which will be ON less (duty cycle decreases), and as a consequence the output voltage drops to 5V.

When the switch is ON, the current will flow from pin8 through the inductor (L1) into Cout. So, the output capacitor gets charged. When the switch is OFF, the energy stored in the inductor, maintains current flow to the capacitor (and load), circulating through the forward biased diode (D1).

Since the input power equals the output power plus the power dissipation in the part, the input current depends on the load current and the input voltage. For low input voltages, the input current will be high but will never exceed the output current, which is the load current.

LM2594 Inverting Regulator



The inverting regulator converts a positive input voltage to a negative output voltage with a common ground. The circuit operates by bootstrapping the regulators ground pin to the negative output voltage, then grounding the feedback pin, the regulator senses the inverted output voltage and regulates it.

This example uses the LM2594 to generate a -5V output. Other output voltages are possible by selecting other output voltage versions, including the adjustable version. Since this regulator topology can produce an output voltage that has a magnitude either greater than or less than the input voltage, the maximum output current greatly depends on both the input and output voltage. The curve shown provides a guide as to the amount of output load current possible for the different input and output voltage conditions.

The maximum voltage appearing across the regulator is the absolute sum of the input and output voltage, and this must be limited to a maximum of 40V. For example, when converting +20V to -12V, the regulator would see 32V between the input pin and ground pin. The LM2594 has a maximum input voltage rating of 40V.

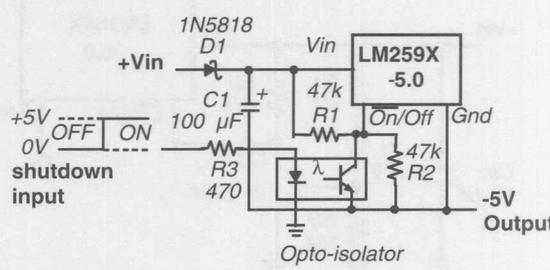
Additional diodes are required in this regulator configuration. Diode D1 is used to isolate input voltage ripple or noise from coupling through the C1 capacitor to the output, under light or no load conditions. Also, this diode isolation changes the topology to closely resemble a buck configuration thus providing good closed loop stability. A Schottky diode is recommended for low input voltages (because of its lower voltage drop), but for higher input voltages, a fast recovery diode could be used.

Without diode D3, when the input voltage is first applied, the charging current of C1 can pull the output positive by several volts for a short period of time. Adding D3 prevents the output from going positive by more than a diode voltage.

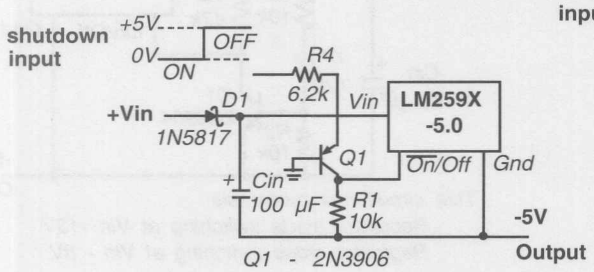
This type of inverting regulator can require relatively large amounts of input current when starting up, even with light loads. Input currents as high as the LM2594 current limit (approx. 0.8A) are needed for at least 2 ms or more, until the output reaches its nominal output voltage. Input power sources that are current limited or sources that can not deliver these currents without getting loaded down, may not work correctly. Because of the relatively high startup currents required by the inverting topology, the delayed startup feature (C1, R1 & R2) shown is recommended. By delaying the regulator startup, the input capacitor is allowed to charge up to a higher voltage before the switcher begins operating. A portion of the high input current needed for startup is now supplied by the input capacitor (C1).

Ground Referenced Shutdown of LM2594/5/6 Inverting Regulators

Logic low is OFF



Logic high is OFF



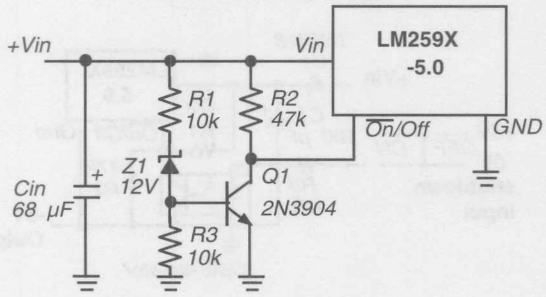
To use the $\overline{\text{ON}}$/OFF control in a standard buck configuration is simple: pull the $\overline{\text{ON}}$/OFF pin below 1.3V (@25°C, referenced to ground) to turn regulator ON, pull it above 1.3V to shut the regulator OFF. With the inverting configuration, some level shifting is required, because the ground pin of the regulator is no longer at ground, but is now at the negative output voltage level. Two different shutdown methods for inverting regulators are shown.

In the left figure, Q1 is not conducting in the ON condition, so the $\overline{\text{ON}}$/OFF pin will be pulled low. In the OFF condition, Q1 is conducting current, so the $\overline{\text{ON}}$/Off pin will be pulled high.

For the right figure, a similar reasoning can be followed, although the logic of the transistor is reversed (OFF, transistor off; ON, transistor on).

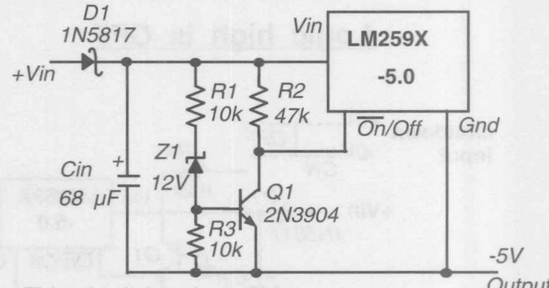
LM2594/5/6 Undervoltage Lockout

Undervoltage lockout for buck regulator



Z1 is a 1N5242B

Undervoltage lockout with hysteresis for inverting regulator



This circuit has hysteresis
Regulator starts switching at Vin -13V
Regulator stops switching at Vin - 8V

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Some applications require the regulator to remain off until the input voltage reaches a predetermined voltage. An undervoltage lockout feature applied to a buck regulator is shown in the right figure; the left figure applies the same feature to an inverting circuit. The circuit in the right figure features a constant threshold voltage for turn on and turn off (zener voltage plus approximately one volt). The inverting regulator circuit in the left figure has a turn ON voltage which is different than the turn OFF voltage. The amount of hysteresis is approximately equal to the value of the output voltage. If zener voltages greater than 25V are used, an additional 47kΩ resistor is needed from the $\overline{\text{ON}}$/Off pin to the ground pin to stay within the 25V maximum limit of the $\overline{\text{ON}}$/OFF pin referred to the Gnd pin..

Features and Benefits LM2594/5/6

- **150 kHz frequency of operation** ⇨ **Smaller inductors and caps for reduced PCB footprint**
- **4 external components** ⇨ **Ease of design, minimum board space**
- **3.3V, 5V, 12V, and adj. vers.** ⇨ **Solutions to any voltage step down application**
- **TTL compatible shutdown** ⇨ **Uncluttered digital control**
- **New "Switchers Made Simple" version 4.2 software** ⇨ **Design power converters in minutes**
- **Multi-sourced standard inductors** ⇨ **Readily available components**
- **Guaranteed system converter specs** ⇨ **Guaranteed converter functionality**

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The LM259X series of step down Simple Switchers has some great features which lead to substantial benefits :

• The **power switch** of this regulator operates at **150 kHz**, which makes it possible to select and use smaller inductors and capacitors. This makes it possible to make smaller Printed Circuit Boards (PCB) because of the reduced footprints of these components.

•There are **only four external components** needed for the basic designs, which makes it easy to design and to use. Also the required board space is minimal.

•There are **3.3V, 5V, 12V and adjustable versions** of the LM259X. So, there will be a solution for any buck regulator application.

•The $\overline{\text{ON}}$/OFF pin is **TTL compatible**, so the digital control is straight forward and easy to use.

•Together with the release of this part, there will be the release of the new **"Switchers Made Simple" version 4.2 software**. This allows you to design a power supply in minutes.

•Because of the **multi-sourced standard inductors**, that are readily available, there won't be a problem of finding the right components.

•As with all Simple Switchers, the system converter specifications are guaranteed, which guarantees the LM259X's functionality.

•The smallest package offered is a **SO-8 surface mount** package for the LM2594. This allows once more for minimal board area.

Overall, this is a product with several great features which makes it attractive to use.

Extra Features and Benefits with LM2597/8/9

• **Internal flag for "power OK" indicator with programmable delay**	• **For μP reset indicator**
• **Out of regulation flag**	• **For fault indicator**
• **Softstart**	• **Limits in-rush current**
• **Bias Supply***	• **Higher efficiency**

***LM2597 only**

The LM2597/8/9 are similar to the LM2594/5/6, but offer a few more interesting features, such as :

•A **internal flag for "power OK" indicator**. When the output voltage reaches 95% of the nominal ouput voltage, this flag goes high, after a delay programmed with an external capacitor. This flag can be used for a μP reset

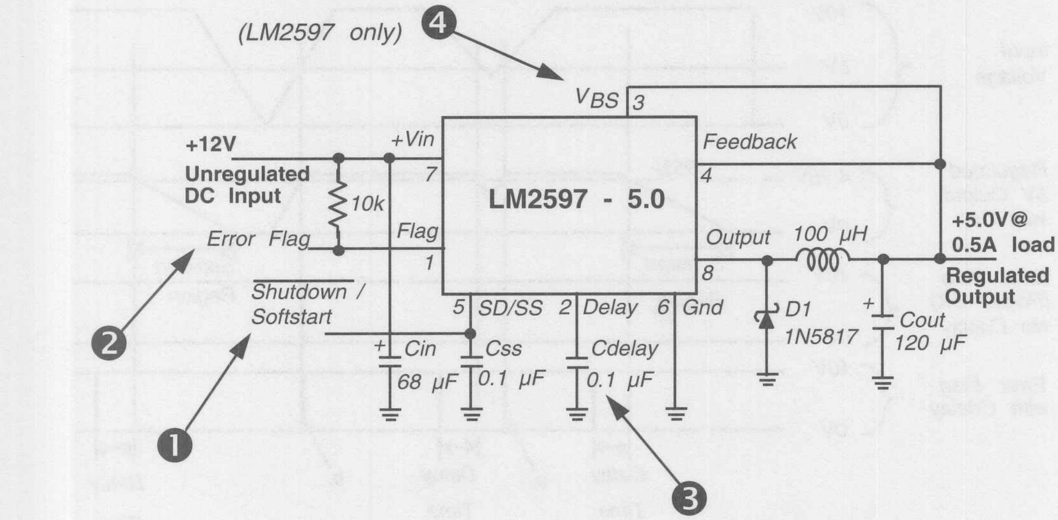
•When the output of the LM2597/8/9 drops below 95% of its output voltage, the **out of regulation flag** (error flag) will indicate a fault condition.

•The **softstart** feature makes sure that the in-rush current is limited, to avoid overloading the source supply during start-up.

•The **Bias Supply** pin can be used for better regulator efficiency, especially at low output currents and high input voltages. This feature is only available on the LM2597.

LM2597/8/9 adds Features to LM2594/5/6

Fixed output voltage buck regulator



The LM2597 is a switching step down regulator, similar to the LM2594, with extra features.

The LM2597 uses 3 extra pins, the Bias Supply pin, the Delay pin and the Error Flag pin, and the ON/OFF pin has been replaced with the Shutdown/Softstart pin.

Shutdown/Softstart (pin 5) - This dual function pin provides the following features:

(a) Allows the switching regulator circuit to be shut down using logic level signals, thus dropping the total input supply current to approximately 80 μA.

(b) Adding a capacitor to this pin provides a softstart feature which minimizes startup current and provides a controlled ramp up of the output voltage.

Error Flag (pin 1) - Open collector output that provides a low signal (flag transistor ON) when the regulated output voltage drops more than 5% from the nominal output voltage. On start up, Error Flag is low until Vout reaches 95% of the nominal output voltage and after a delay time, determined by the Delay pin capacitor. This signal can be used as a reset to a microprocessor on power-up.

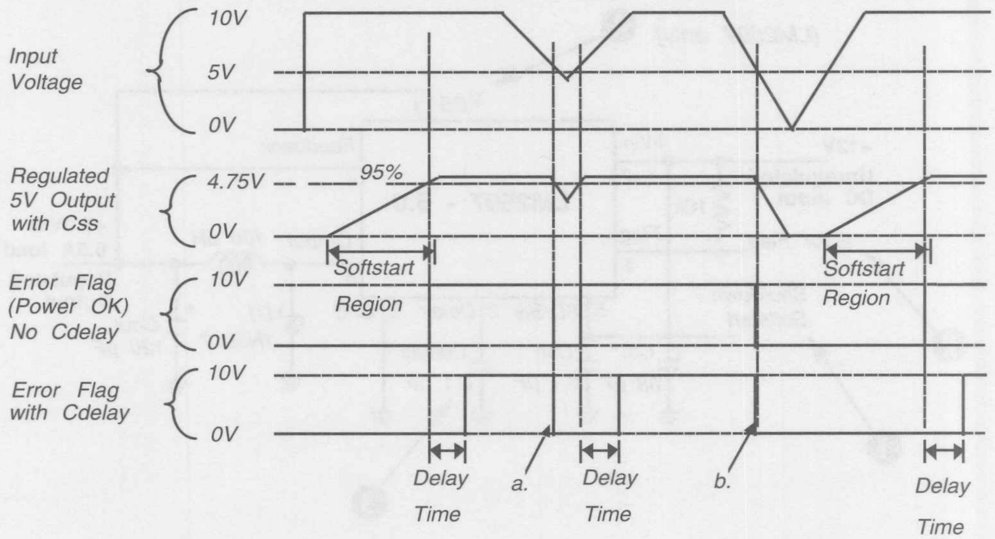
Delay (pin 2) - At power-up, this pin can be used to provides a time delay between the time the regulated output voltage reaches 95% of the nominal output voltage, and the time the error flag output goes high.

Bias Supply (pin 3) - This feature allows the regulator's internal circuitry to be powered from the regulated output voltage or an external supply, instead of the input voltage. This results in increased efficiency under some operating conditions, such as low output current and/or high input voltage. Again, this feature is only available with the LM2597.

Special Note If any of the above four features (Shutdown/Softstart, Error Flag, Delay, or Bias Supply) are not used, the respective pins should be left open.

LM2597/8/9 Error Flag with Delay

Timing diagram for 5V output



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The delay capacitor provides a delay for the error flag output (see timing diagram). A capacitor on this pin provides a time delay between the time the regulated output voltage (when it is increasing in value) reaches 95% of the nominal output voltage, and the time the error flag output goes high.

A 2.5μA constant current from the delay pin charges the delay capacitor resulting in a voltage ramp. When this voltage reaches a threshold of approximately 1.3V, the open collector error flag output (or power OK) goes high. This signal can be used to indicate that the regulated output has reached the correct voltage and and has stabilized. If, for any reason, the regulated output voltage drops by 5% or more (see point a. and b. in the diagram), the error output flag (pin 1) immediately goes low (internal transistor turns on).

The delay time can be determined as follows : $dt = C \times dV / I = 0.1\mu F \times 1.3V / 2.5\mu A = 52msec$ (or 520msec per μF).

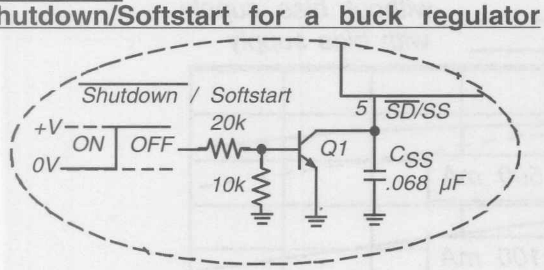
The delay capacitor provides very little delay if the regulated output is dropping out of regulation. The delay time for an output that is decreasing is approximately a 1000 times less than the delay for the rising output. For a 0.1mF delay capacitor, the delay time would be approximately 50 msec when the output is rising and passes through the 95% threshold, but the delay for the output dropping would only be approximately 50 μsec.

The error flag output (or power OK), is the collector of a NPN transistor, with the emitter internally grounded. To use the error flag, a pull-up resistor to a positive voltage is needed. The error flag transistor is rated up to a maximum of 45V and can sink approximately 3 mA.

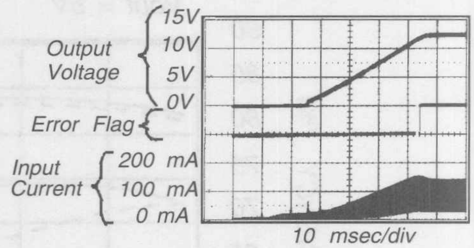
If the error flag is not used, it can be left open.

LM2597/8/9 $\overline{\text{Shutdown}}$ / Softstart

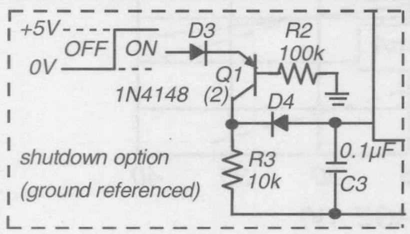
$\overline{\text{Shutdown}}$/Softstart for a buck regulator



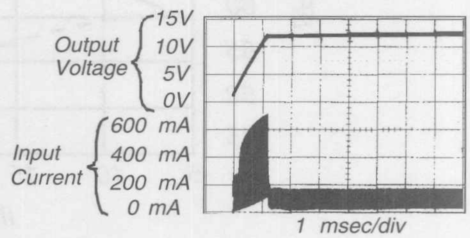
With Softstart



$\overline{\text{Shutdown}}$/Softstart for an inverting regulator



Without Softstart



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The circuit shown in the upper left corner is a typical $\overline{\text{Shutdown}}$/Softstart circuit for a standard buck regulator. The photos on the right hand side show the effects of softstart on the output voltage and the input current for the LM2597, with and without a softstart capacitor. The upper photo also shows the error flag output going high when the output voltage reaches 95% of the nominal output voltage. The reduced input current required at startup is very evident when comparing the two photos. The softstart feature reduces the startup current from 700 mA down to 160 mA, and delays and slows down the output voltage rise time.

This reduction in start up current is useful in situations where the input power source is limited in the amount of current it can deliver. In some applications softstart can be used to replace undervoltage lockout or delayed startup functions.

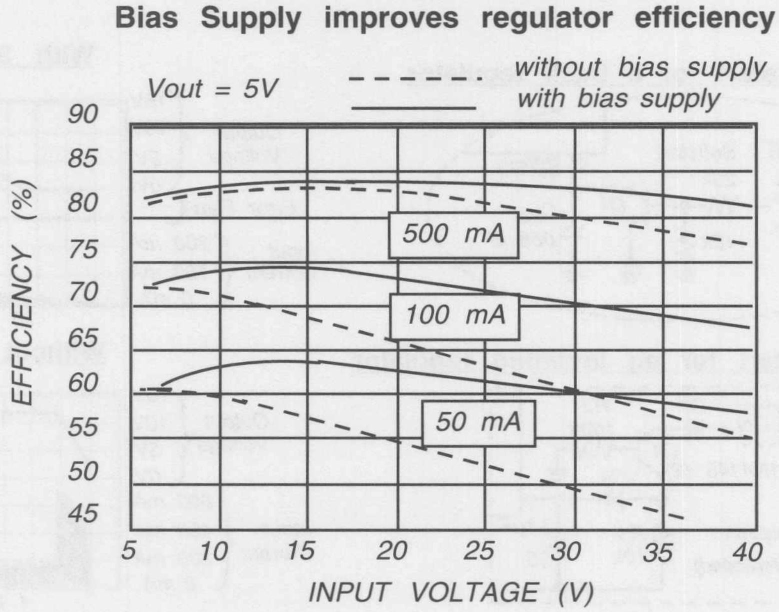
If a very slow output voltage ramp is desired, the softstart capacitor can be made much larger. Many seconds or even minutes are possible. To get the desired time, you can use :
dt = C x dV / I, e.g. dt = 0.068µF x (2.8 - 1.8V) / 1.6µA = 42.5ms

Note that the charge current from $\overline{\text{SD}}$/SS is 1.6µA during the time the output voltage comes up, the softstart region, and the $\overline{\text{SD}}$/SS pin voltage changes from 1.8V to 2.8V. These numbers are fixed, the only variable is the softstart capacitor. The softstart time can be estimated at 600ms per µF (1V x C / 1.6µA = 0.6 sec per µF).

If only the shutdown feature is needed, the softstart capacitor can be eliminated.

Also shown in the circuit at the bottom left is a shutdown method for the inverting configuration. With the inverting configuration, some level shifting is required, because the ground pin of the regulator is no longer at ground, but is now at the negative output voltage. The shutdown method shown accepts ground-referenced shutdown signals.

LM2597 Bias Supply



Bias Supply for improved efficiency

The bias supply (VBS) pin allows the LM2597's internal circuitry to be powered from a power source, other than Vin. Typically the output voltage is used. This feature can increase efficiency and reduce junction temperatures under some operating conditions. The greatest increase in efficiency occurs with light load currents, high input voltage and low output voltage (4V to 12V). In the efficiency curve the solid lines are with the VBS pin connected to the regulated output voltage, while the dashed lines are with the BS pin open.

The bias supply pin requires a minimum of approximately 3.5V at room temperature (4V @-40°C), and can be as high as 30V, but there is little advantage of using the bias supply feature with voltages greater than 15 or 20 Volts. The current required for the Vin pin is typically 4 mA.

To use the bias supply feature with output voltages between 4V and 15V, wire the bias pin to the regulated output. Since the VBS pin requires a minimum of 4V to operate, the 3.3V part can not be used this way. When the VBS pin is left open, the internal regulator circuitry is powered from the input voltage.

LM2594/97 versus LM2574

LM2594/97	LM2574
• 150 kHz frequency of operation	• 52kHz frequency of operation
• SO-8 surface mount pkg	• DIP-8 through hole pkg
• Input voltage range up to 40V	• Input voltage range up to 60V
• Better efficiency with Bias Supply	• Efficiency approximately 85%
• Internal flag for "power OK" indicator with programmable delay	
• Out of regulation flag	
• Softstart	

The new Simple Switcher buck regulators, the LM259X, have been improved compared to the original Simple Switcher buck regulator, the LM2574/5/6.

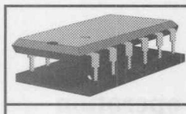
Tripling the frequency of operation allows us to use smaller external components, which makes it possible for smaller PCB design layouts. This is also true for the SO-8 surface mount package compared to the DIP-8 through hole. The new design also provides higher efficiency (under most operating conditions) when the Bias Supply is used.

The LM257X has an input range up to 60V, which is a benefit compared to the 40V limit of the new buck regulators.

The LM25978/9 has several more features compared to the LM2574, and the benefits have been discussed earlier. The LM259X offer more design flexibility compared to the LM257X.

The new Simple Switchers have been improved a great deal compared to the original Simple Switchers.

Packages and Voltage Options



	Max. Output Current	Voltages	Packages
LM2594	500mA	3.3, 5, 12, ADJ	N-8, M-8
LM2597	500mA	3.3, 5, 12, ADJ	N-8, M-8
LM2595	1A	3.3, 5, 12, ADJ	TO220-5, TO263-5
LM2598	1A	3.3, 5, 12, ADJ	TO220-7, TO263-7
LM2596	3A	3.3, 5, 12, ADJ	TO220-5, TO263-5
LM2599	3A	3.3, 5, 12, ADJ	TO220-7, TO263-7

66

All of the new SIMPLE SWITCHER buck regulators are available in 3.3V, 5V, 12V and adjustable (ADJ) output versions.

Both 0.5A converters are available in 8-pin surface-mount (M) and 8-pin dual-in-line (N) packages. The higher-current converters are available in the TO-220 (T) through-hole power package and the TO-263 (S) surface-mount package. The basic 1A and 3A converters are in 5-pin packages, and the full-featured 1A and 3A versions are in 7-pin packages.

The full part numbers indicate the product family, package, and voltage option. For example, the order numbers for the **LM2594** (basic 0.5A converter) are :
LM2594N-3.3, LM2594N-5.0, LM2594N-12, LM2594N-ADJ for through-hole
and
LM2594M-3.3, LM2594M-5.0, LM2594M-12, LM2594M-ADJ for surface-mount.

The order numbers for the **LM2599** (full-featured 3A converter) are :
LM2599T-3.3, LM2599TN-5.0, LM2599T-12, LM2599T-ADJ for through-hole
and
LM2599S-3.3, LM2599S-5.0, LM2599S-12, LM2599S-ADJ for surface-mount.

Part numbers for the other new SIMPLE SWITCHERs would be constructed similarly.

Pinout Compatibility: Check It First!

- **Basic New converters vs. Original converters**
  - **New, Original 0.5A parts have *different* pinouts**
  - **New, Original 1A parts have Vin & Vout *reversed***
  - **New, Original 3A parts have same pinouts**
- **Basic vs. Full-Featured New converters**
  - **0.5A parts have similar pinouts, *reversed* logic**
  - **1A parts have same pin 1 & 2, others different**
  - **3A parts have same pin 1 & 2, others different**

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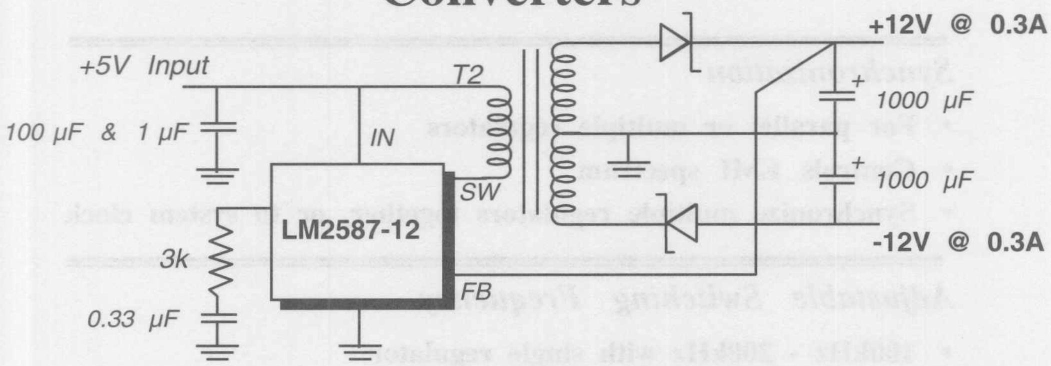
LM2585, LM2586 LM2587, LM2588

New SIMPLE SWITCHER™ Flyback and Boost Regulators

LM2585/6/7/8: New Flyback Converters



- **3.3V, 5V, 12V, and ADJ**
- **5A/65V switch (LM2587/8); 100kHz oscillator**
- **Delivers over 25W (Over 50W in some applications)**
- **ON/OFF Control (LM2586/8)**

More new products in the SIMPLE SWITCHER family are flyback regulators, LM2585, LM2586, LM2587 and LM2588. As with other SIMPLE SWITCHER converters, a complete DC/DC converter design requires only a few external components, which are well-specified in the product data sheet.

In the flyback regulator configuration (shown above), the LM2587 can control output voltages of 3.3V, 5V, or 12V with the fixed voltage versions, or a user-defined output voltage with the "ADJ" version. The input voltage range is from 4V to 40V.

The internal switch of the LM2587 is driven by a 100 kHz oscillator, which allows the use of smaller components (compared with lower switching frequencies). The switch can stand off 65V, and is rated for 5A peak. The LM2585 is rated at 3A peak.

The LM2588 (LM2586) has the same ratings and performance as the LM2587 (LM2585), but has the additional options of ON/OFF control and frequency sync/adjust.

The LM2585/7 is available in the 5-lead TO-220 package (with bent, staggered leads), and the new surface-mount 5-lead TO-263 package. The LM2586/8 is available in similar 7-pin TO-220 and TO-263 packages.

LM2586/8: Control Switching Frequency

Synchronization

- For parallel or multiple regulators
- Controls EMI spectrum
- Synchronize multiple regulators together, or to system clock

Adjustable Switching Frequency

- 100kHz - 200kHz with single regulator
- Adjustment with one resistor
- Smaller designs at high frequency

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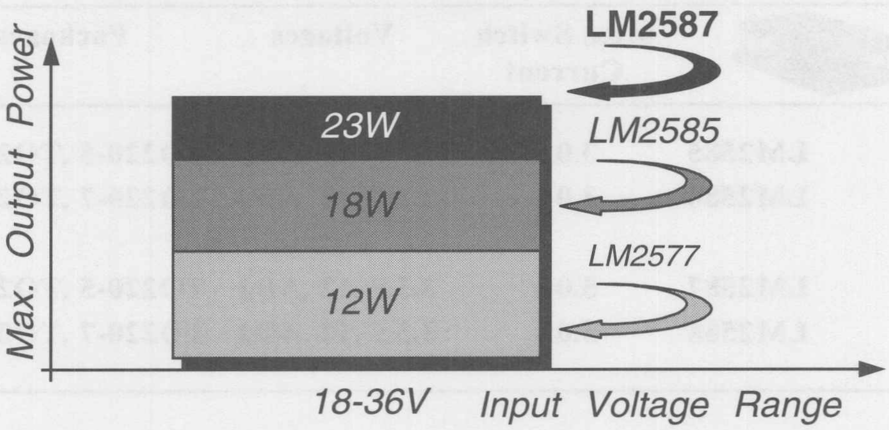
While the LM2585 and LM2587 operate at a fixed switching frequency of 100 kHz, the LM2586 and LM2588 allows its switching frequency to be modified. Using either of two control methods, the switching frequency can be varied between 100 kHz (its base frequency) to 200 kHz.

Synchronization forces the LM2586/8 switching frequency to match that of an external source, such as another switching regulator or a system clock. This keeps the EMI generated within the system to a predictable set of frequencies. It also prevents similar (but not matching) switching frequencies from producing a beat frequency. Both of these results make it easier to filter out switching noise in the system.

The second frequency control method changes the switching frequency of an LM2586/8, using a single resistor (from the Frequency Adjust pin to ground). This allows a design to be customized for a smaller size, as the sizes of the transformer and output capacitor tend to go down as the switching frequency increases.

LM2587's 5A Switch Delivers 92% More Power than LM2577



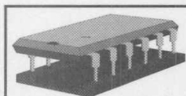
Designs based on same size transformer, 12V output

The LM2585, LM2586, LM2587, and LM2588 are similar to the original SIMPLE SWITCHER LM2577, but with additional capabilities. While the LM2577 can deliver about 15W with its 3A switch, the 5A LM2587/LM2588 is able to deliver up to about 23W of output power. (In selected applications, the LM2587/LM2588 can deliver over 50W to a load.) In addition, the LM2587 switching frequency is roughly twice that of the original products, which helps minimize the size of the external components.

As a SIMPLE SWITCHER converter, the LM2585/6/7/8 share several features with the LM2577. The regulated output voltage of an LM2585/6/7/8 DC/DC converter is guaranteed to have a tolerance of better than 5% over its full line, load, and temperature ranges (4% at room temperature). The internal switch is protected by current limit and over-temperature shutdown.

As in the LM2577, the regulator control method used in the LM2585/6/7/8 is current-mode control, with a soft-start function which minimizes inrush current during startup.

Packages and Voltage Options

	Max. Switch Current	Voltages	Packages
LM2585	3.0A	3.3,5 ,12 ,ADJ	TO220-5 ,TO263-5
LM2586	3.0A	3.3,5 ,12 ,ADJ	TO220-7 ,TO263-7
LM2587	5.0A	3.3,5 ,12 ,ADJ	TO220-5 ,TO263-5
LM2588	5.0A	3.3,5 ,12 ,ADJ	TO220-7 ,TO263-7

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All of the new SIMPLE SWITCHER flyback / boost regulators are available in 3.3V, 5V, 12V and adjustable (ADJ) output versions.

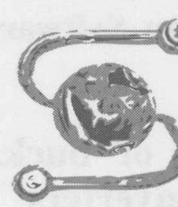
Both the LM2585 (3Amp) and LM2587 (5Amp) converters are available in the TO-220 (T) through-hole power package and the TO-263 (S) surface-mount package. The basic 3A and 5A converters (LM2585 and LM2587) are in 5-pin packages, and the full-featured 3A and 5A versions (LM2586 and LM2588) are in 7-pin packages.

The full part numbers indicate the product family, package, and voltage option. For example, the order numbers for the **LM2588** (full-featured 5A converter) are :

LM2588T-3.3, LM2588T-5.0, LM2588T-12, LM2588T-ADJ for through-hole
and
LM2588S-3.3, LM2588S-5.0, LM2588S-12, LM2588S-ADJ for surface-mount.

Part numbers for the other new SIMPLE SWITCHERs would be constructed similarly.

National Semiconductor on the WEB

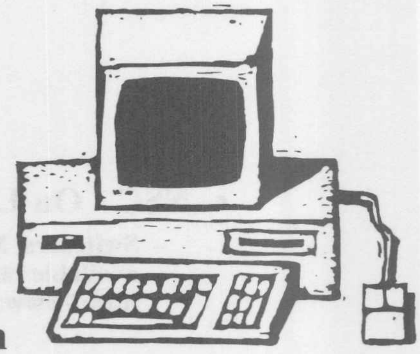


- **NSC On-Line (http://www.national.com)**
  - **Switchers Made Simple DC/DC Converter design software is available at http://www.national.com/news/switch/switch.htm**

FOR DESIGN ENGINEERS.

Switchers Made Simple (Version 4.2)

Power Supply Design Software



- **Automated design of buck, flyback, boost converters**
- **Companion to new SIMPLE SWITCHER converters**
- **New menus, design flow for even faster power supply design**

To go along with the newest SIMPLE SWITCHER regulators, there is a new version of the SIMPLE SWITCHER design software, "Switchers Made Simple" (Ver. 4.2). This software provides for the automated design of step-down, flyback and boost converters, and is a companion to the new LM259X and LM2585/6/7/8.

Similar in concept to the original "Switchers Made Simple" (which was developed for the original SIMPLE SWITCHER products, and is included with Ver. 4.1), the new version has a new design flow for almost-instant design of DC/DC converters. In addition, new menus make modification of the design a matter of a few keystrokes.

For software which supports the LM2574/5/6 and LM2577, please request SMS V4.1 or 3.3.

Switchers Made Simple Available for all SIMPLE Products

SMS V4.2
Buck, Boost, Flyback with New SIMPLE SWITCHER LM2594/5/6/7/8/9 LM2585/6/7/8

- **V4.2 does not include V3.3.**

- **For designs using the original SIMPLE SWITCHERs, download V4.1 which includes V3.3**

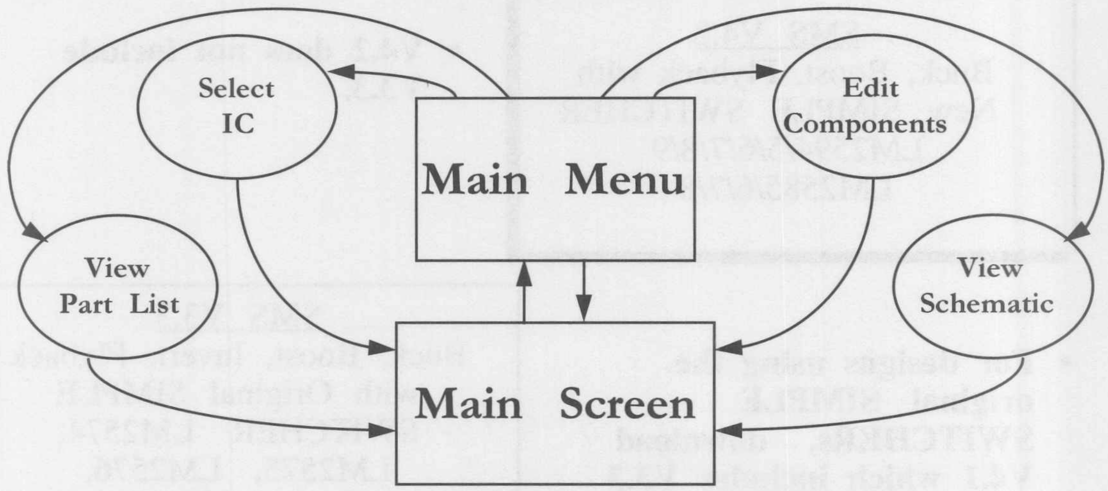
SMS V3.3
Buck, Boost, Invert, Flyback with Original SIMPLE SWITCHER LM2574, LM2575, LM2576, LM2577

In general, Version 4.2 should be used to try any new design. If the design requires one of the original SIMPLE SWITCHERs, the user will be prompted to move to Version 3.3. This will occur primarily for step down designs with input voltages over 40V. The original SIMPLE SWITCHER step down products include "HV" versions with input voltages up to 60V.

To assure support of the original SIMPLE SWITCHER products, their design software "Switchers Made Simple" Version 3.3 is included with the Version 4.1. A common "shell" connects the two parts of the software together, and allows the user to switch from one to the other without completely exiting to DOS.

In addition, the shell includes a brief description of both versions of the software. This allows the user to understand the capabilities of each version.

Switchers Made Simple (Version 4.2) Flow Chart



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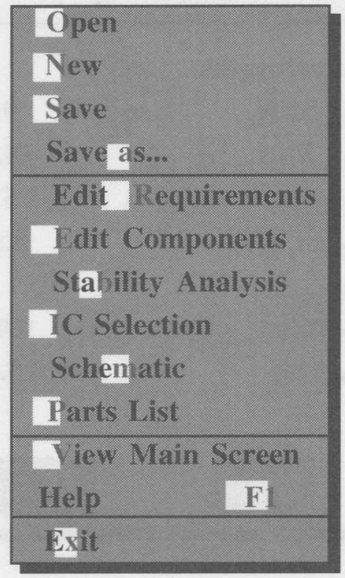
"Switchers Made Simple" (Ver. 4.2) is a menu driven program. The Main Menu, at the center of the flow chart, allows the user to easily maneuver through the program, either to design a new power supply or to modify an existing one. The power supply specifications, component values, and some results of calculations (such as peak switch current) are displayed on the Main Screen.

From the Main Menu, a schematic or the part lists can be viewed and printed. Once the user exits from these screens the program returns the user to the Main Screen. The user can go to the Main Menu by pressing the "esc" (escape) key.

Main Menu



"Hot" keys allow fast editing or file operations without having to call up this Main Menu.

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This is the Main Menu screen. From here the user can do the following:

- Open an old design (Open)
- Create a new design (New)
- Save a design (Save)
- Save a design to a different name (Save as)
- Edit the user defined inputs (Edit Requirements)
- Edit component values (Edit Components)
- View the cross-over frequency and phase margin values (Stability Analysis)
- Select an alternate National Semiconductor IC (IC selection)
- View the schematic (Schematic)
- View the parts list (Parts List)
- View the operating conditions (View Main Screen)
- Ask for help (Help)
- Exit program (Exit)

Every one of the menu item has a "Hot Key" associated with it. For example, the "P" is the hot key for displaying the parts list. The hot keys allows the user to maneuver through the program quickly, without the requirement to first call up the Main Menu.

Input Menu Defines Design

NumOutputs	**=**	**1**
V_{IN} Min	**=**	**9.00 V**
V_{IN} Max	**=**	**18.00 V**
$V_{OUT}1$	**=**	**5.00 V**
$I_{OUT}1$ Max	**=**	**0.50 A**
$V_{ripple}1$	**=**	
T_A Min	**=**	**0.00 °C**
T_A Max	**=**	**70.00°C**

Ok Cancel

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This is the design specification screen, where the user inputs the design requirements. This screen provides a template in which the user specifies the input voltage range, output voltage and current, output ripple voltage, minimum ambient temperature, and maximum ambient temperature. The user proceeds to the next screen by selecting the "Ok" button or by pressing the "O" hot key.

In contrast with the previous versions of "Switchers Made Simple," the user does not need to specify the type of power supply to be designed. With Version 4.2, the software automatically determines the power supply type. In this release, a step-down, a flyback or a boost design is chosen, matching the capabilities of the new SIMPLE SWITCHER products.

If the application requires a higher voltage (>40V input) step-down supply or an inverting power supply design, the user is instructed to use the previous version of the software, Ver. 3.3, which is included as part of Ver. 4.1.

choose 5.0001 TO FORCE USE OF ADJUSTABLE PARTS.

Main Screen Summarizes Design

Input Requirements		Operating Values*		Component Values	
V_{IN} Min	= 9.00 V	Frequency	= 150.00 kHz	L	= 150.00 uH
V_{IN} Max	= 18.00 V	Duty Cycle	= 65.68 %	L DCR	= 0.11Ohms
$V_{OUT}1$	= 5.00 V	IC I_{PK} Max	= 0.65 A	C_{IN}	= 47.00 uF
$I_{OUT}1$ Max	= 0.50 A	IC I_{PK}	= 0.54 A	C_{IN} ESR	= 0.34Ohms
T_A Min	= 0.00 °C	L I_{PP}	= 84.74 mA	C_{OUT}	= 56.00 uF
T_A Max	= 70.00 °C	Efficiency	= 81.76 %	C_{OUT} ESR	= 46.00 m~~W~~ Ω
		I_C P_D	= 0.421 W	ESR AC	= 0.20~~Ohms~~ mΩ
		I_C T_J	= 106.54 °C	Cu Area	= 1.50 in^2
		Diode P_D	= 0.25 W	Cu Wgt	= 1.00 oz
		L P_D	= 0.46 W		
		Cross Freq	= 10.99 kHz		
		Phase Marg	= 22.18 Deg		
		V_{OUT} p-p	= 23.73 V		

max value stability calc.

* *All values not shown due to presentation constraints*

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The main screen contains a detailed view of the converter design and operating conditions. From here the user can determine if the converter performance meets the design requirements. More importantly, this screen allows the user to view the effects on the converter operating condition when alternate component values are chosen.

Key parameters for the IC include peak switch current (IC Ipk) and total power dissipated (IC Pd). The recent trend toward smaller packages makes proper thermal design very important. "Switchers Made Simple" will determine if a heatsink is required. If a heatsink is required, "Switchers Made Simple" will calculate the heatsink thermal resistance using the IC's power dissipation and junction to case thermal resistance.

"Switchers Made Simple" (Ver. 4.2) always defaults to a through-hole package. If a surface mount package can be used in the application, the program will list it in the IC Selection screen. Simply select the desired part from this screen, then press the "O" hot key. The program will automatically calculate the printed circuit board copper area and weight to keep the IC's maximum junction temperature below 110°C.

"Switchers Made Simple" calculates all operating conditions based on the worst-case input voltage and maximum output current. For flyback and boost converters, the worst-case input voltage is the minimum input. For step-down converters, the maximum input is worst-case. However, the other extreme of the specified input voltage range is also checked, to be sure the product will be operating within its ratings.

New SIMPLE SWITCHER Converters

	Output Power/Current	Switching Frequency	Product Features
Flyback			
–LM2585	≥ 15W	100 kHz	
–LM2586	≥ 15W	100 kHz	Shutdown; Sync; Freq. Adj.
–LM2587	≥ 25W	100 kHz	
–LM2588	≥ 25W	100 kHz	Shutdown; Sync; Freq. Adj.
Step-Down (Buck)			
–LM2594	0.5A	150 kHz	Shutdown
–LM2595	1.0A	150 kHz	Shutdown
–LM2596	3.0A	150 kHz	Shutdown
–LM2597	0.5A	150 kHz	Shutdown; Softstart; V-Bias; Reset
–LM2598	1.0A	150 kHz	Shutdown; Softstart; Reset
–LM2599	3.0A	150 kHz	Shutdown; Softstart; Reset

... all supported by "Switchers Made Simple" v4.2

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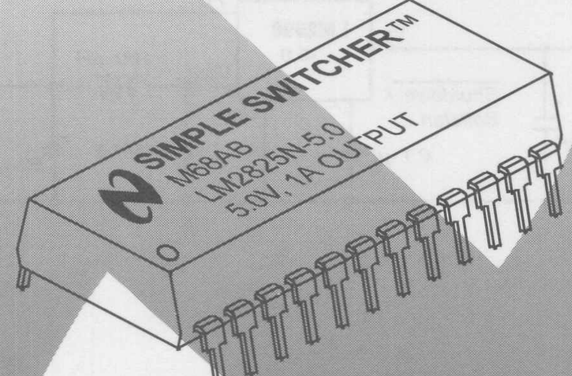
The new SIMPLE SWITCHERs shown in this presentation are noted above. These products extend the capability of the SIMPLE SWITCHER family, to higher output power, higher switching frequency, and additional control features.

ORIGINAL SIMPLE SWITCHER DC/DC Converter Products

Part Number	Output power/	Switching current	Features frequency
Flyback			
LM2577	15W	52kHz	
Step-Down			
LM2574	0.5A	52kHz	Shutdown
LM2575	1A	52kHz	Shutdown
LM2576	3A	52kHz	Shutdown

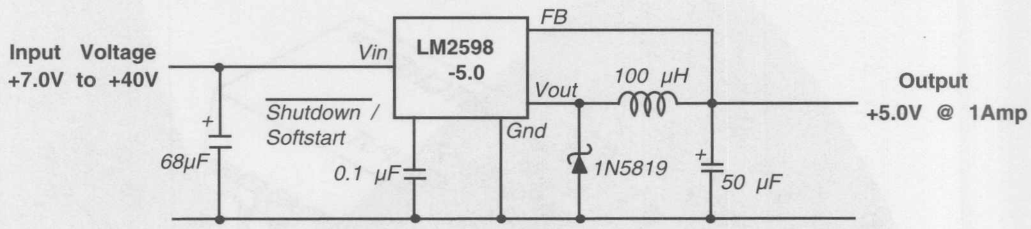
These parts are supported by SMS V3.3.

LM2825
DC/DC Converter IC



The New SIMPLE SWITCHER™
The Simplest Switcher

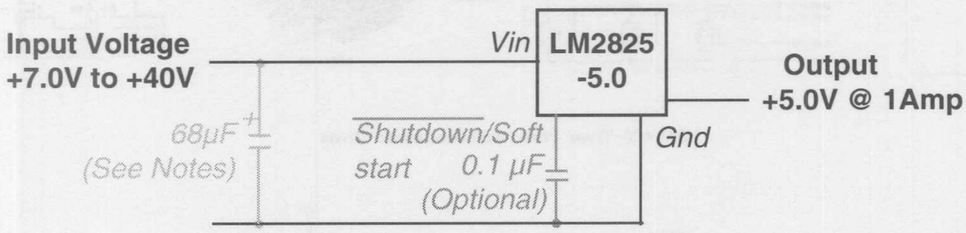
Simple Switcher



The new LM2598 Simple Switcher™ provides an easy to use solution for step down switching voltage regualtor. It includes an ON/OFF function and soft start capability which limits startup current. It requires only 4 components a diode, inductor, output capacitor, and input capacitor. The softstart capacitor is optional.

For those desiring the ultimate in simplicity a new breakthrough product is on the next page.

Simplest Switcher



- **IC Level Reliability!**
- **No External Components Required!**
- **Available in 3.3V and 5.0V versions**

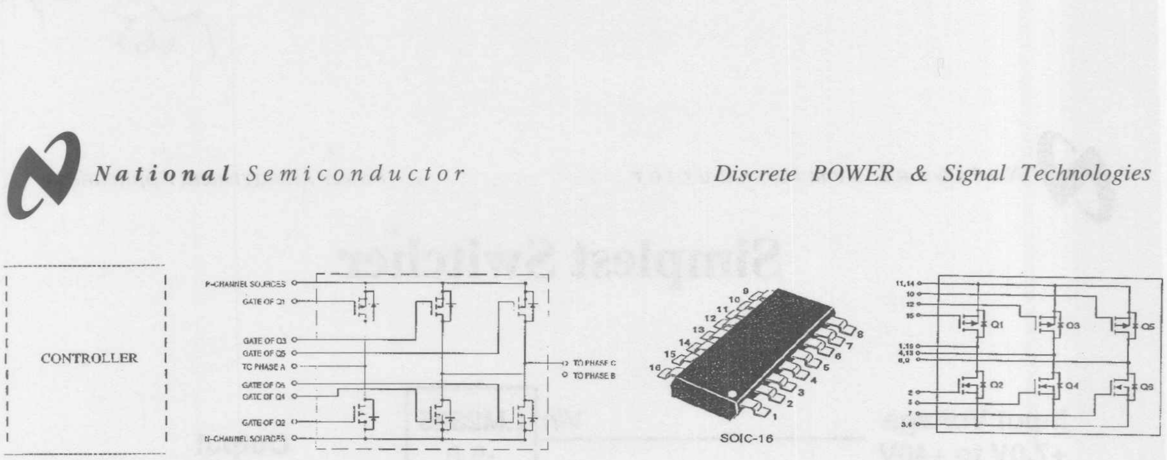
The LM2825 DC/DC Converter IC is the newest member of the Simple Switcher™ product family. IT REQUIRES NO EXTERNAL COMPONENTS.

The diode, output capacitor, inductor, and input capacitor have been incorporated into a single 24 pin DIP. As a result, you now only need to purchase a single device. It provides a full amp of output current up to 70°C. It is currently available in 3.3V and 5.0V versions, with 12V and Adjustable versions becoming available this year.

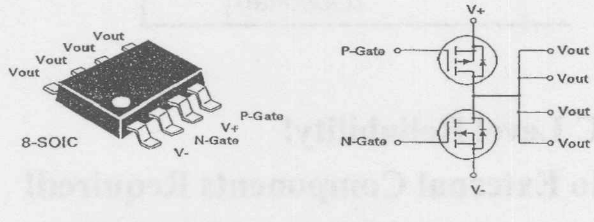
This new technology provides IC level reliability, far exceeding reliability levels of conventional modules as well as conventional board level designs. Radiated EMI testing has been performed, and emissions exceed CISPR 22 Class B requirements.

Features include TTL shutdown, thermal shutdown, current limit, <40mV output ripple in most applications, and soft start (using an additional capacitor) to limit start-up current surges.

Although a bypass input capacitor is incorporated into the LM2825, an additional input capacitor (68uF shown above) is required if leads exceed 6 inches in length from the main power supply or other bypass capacitors. It is also recommended that a 0.1uF softstart capacitor be used in high ambient temperature conditions (>60°C) where the unit is being turned on into a full load.



NDM3000 Three Phase Brushless Motor Driver



NDS8858H Complementary MOSFET Half Bridge

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BRUSHLESS MOTOR CONTROL

Power BJT's as Brushless motor Drivers, have not been the designers choice. Power bipolars are not favored because they can not be driven directly from an IC controller. Power MOSFET's seem to be the ideal choice since they are easy to drive, efficient and cost affective. The power module NDM3000 is a three phase bridge housed in a sixteen pin package. It's low Rds - ON and high break down voltage are the attributes which designers like most.

If higher drive currents are the primary concerns then three NDS8858H (complementary half bridge) can be selected for the design.

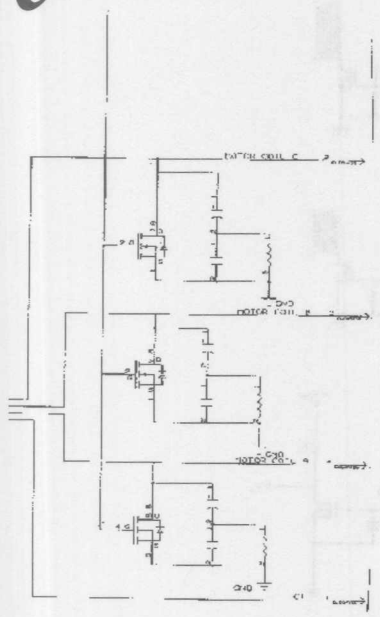


Figure - Brushless Spindle Motor Breaking Circuit

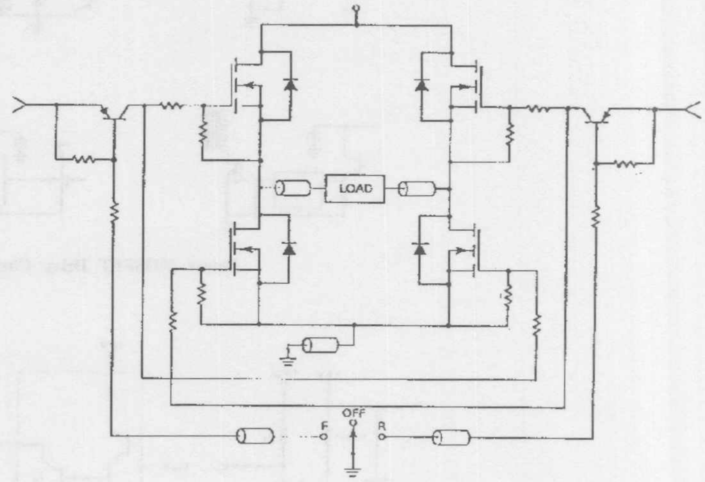


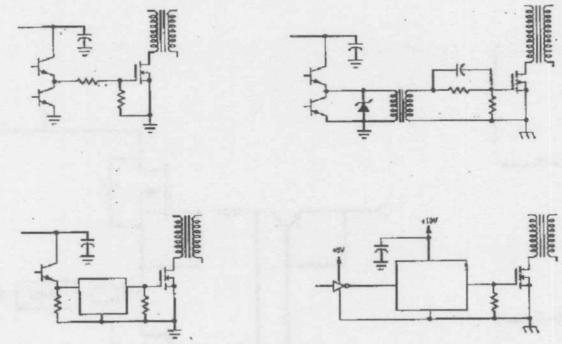
Figure - VCM driver circuit

Brushless Spindle Motor Breaking Application

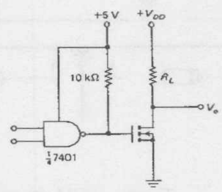
A typical disc drive spindle motor breaking circuit is shown on the left. NDS9956A and NDS9952A are most suitable devices for this application.

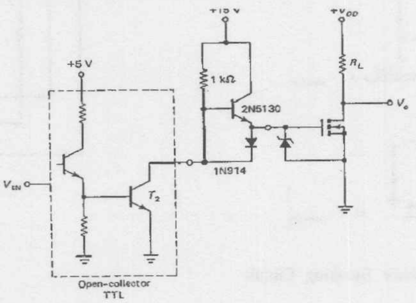
Power Drivers for Disk Drive Head Positioning VCM

A typical Power MOSFET driver circuit for a VCM is shown on the right. NDS8852H and NDS8858H are the Complementary MOSFET Half Bridge packages. Their low Rds - ON and fast switching time make them well suited for this application.



Power MOSFET Drive Circuits





Power MOSFETs as switches

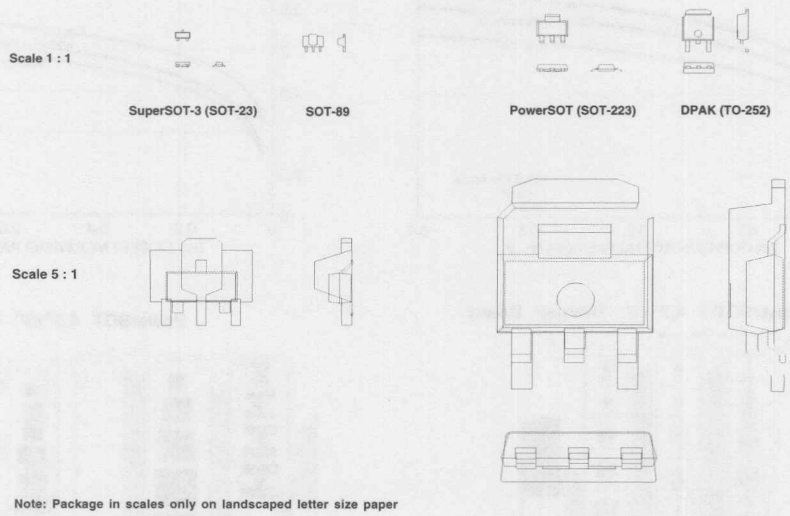
Power MOSFETs are becoming more popular as power switches in switching power supplies . Their main advantages over BJT are such as switching five to ten times faster than BJT and being easier to drive .

These are voltage driven devices and can be driven from controller ICs which have totem pole output drivers with less than 100 ns switching time. Following figure shows some of the common gate drivers .

TTL as Power MOSFET driver

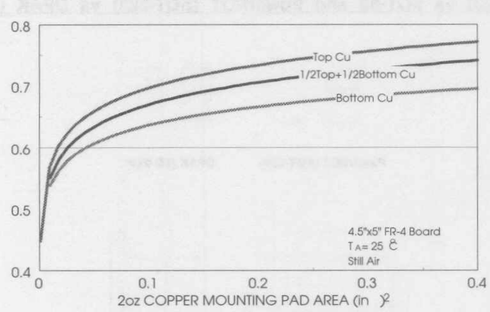
Open collector TTL with a pull up resistor tied to a higher voltage than +5V is an ideal driver for the power MOSFET. The turn ON time is limited by the RC time constant of pull-up and Cin . It offers an excellent current sink to turn OFF power MOSFET quickly .

PACKAGE COMPARISON OF SuperSOT-3 (SOT-23) vs SOT-89 and PowerSOT (SOT-223) vs DPAK (TO-252)

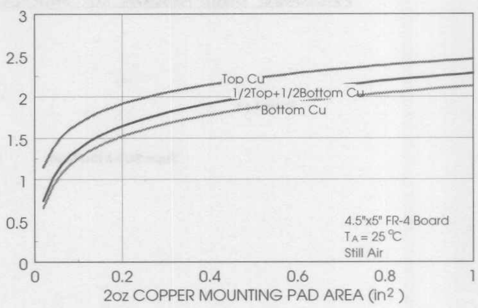


The package footprints of SuperSOT-3 (SOT-23) and PowerSOT (SOT-223) surface mount Power MOSFETs have been miniaturized without sacrificing significant electrical and thermal performance of their counterparts SOT-89 and DPAK (TO-252) respectively. The cost and space differentials truly make these packages superior than their predecessors.

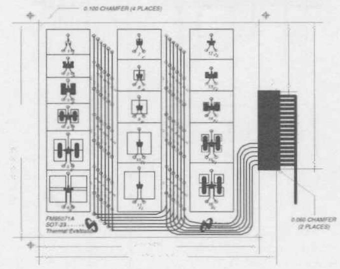
SuperSOT-3 (SOT-23) Power



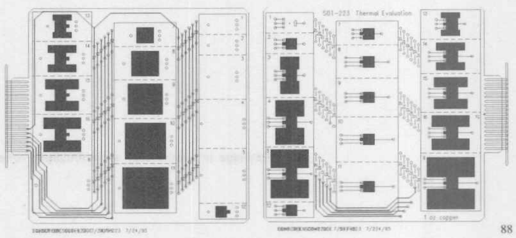
PowerSOT (SOT-223) Power



SuperSOT-3 4.5"x5" Thermal Board



PowerSOT 4.5"x5" Thermal Board



The steady-state power handling capability of SuperSOT-3 (SOT-23) and PoweSOT (SOT-223) are measured by using the thermal boards shown above. Thermal performance is very specific to the applications due to numerous variables; nevertheless, National attempts to provide a general thermal data with devices mounted on a 4.5"x5" FR-4 board with various sizes of top, bottom, and 1/2top+1/2bottom 2oz copper. Measurement is taking under a 1 cubic ft still air at room temperature environment. If board space is limited, layout 2 of package sized copper is recommended to achieve 0.6W for SuperSOT-3 and 1.3W for PowerSOT. If performance is the prime objective over space and cost, increasing the size of the copper can be the most practical thermal solution.

Complete Line of Standard Gate Drive Surface Mount Power MOSFET

	TYPE	SuperSOT™ -3 SOT-23	SuperSOT™ -6 SOT-6	SuperSOT™ -8 SOT-8	SOIC-8 SO-8	PowerSOT SOT-223
PART NUMBER	N P	NDS351N / NDS355N NDS352P / NDS356P	NDC651N NDC652P	NDH8436 NDH8447	SINGLE / DUAL NDS8410 / NDS8936 NDS8435 / NDS8947	NDT455N NDT456P
POWER DISSIPATION		0.5W	1.6W	1.8W	SINGLE / DUAL 2.5W / 2.0W	3.0W
$R_{DS(ON)}$	N P	250mΩ / 125mΩ 500mΩ / 300mΩ	90mΩ 180mΩ	45mΩ 95mΩ	SINGLE / DUAL 15mΩ / 37mΩ 28mΩ / 65mΩ	15mΩ 35mΩ
BV	N P	30V 20V	30V 30V	30V 30V	SINGLE / DUAL 30V 30V	30V 30V
I_D	N P	1.1A / 1.6A 0.85A / 1.1A	3.2A 2.4A	5.8A 4.4A	SINGLE / DUAL 10.0A / 5.3A 7.0A / 4.0A	11.5A 7.3A
TARGET MARKETS		Notebook Cellular Phone PDA Pager	Notebook Cellular Phone PDA PCMCIA	Notebook Cellular Phone HDD PDA	Notebook Cellular Phone HDD	HDD Computers Automotive
TECHNOLOGY	5M cell/in²	4.5V Gate Drive	4.5V Gate Drive	4.5V Gate Drive	10V Gate Drive	10V Gate Drive

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Complete Line of Low Gate Drive Surface Mount Power MOSFET

	TYPE	SuperSOT™ -3 SOT-23	SuperSOT™ -6 SOT-6	SuperSOT™ -8 SOT-8	SOIC-8 SO-8
PART NUMBER	N P	NDS331N / NDS335N NDS332P / NDS336P	NDS631N NDS632P	NDS831N NDS832P	SINGLE / DUAL NDS8426 / NDS8926 NDS8434 / NDS8934
POWER DISSIPATION		0.5W	1.6W	1.8W	SINGLE / DUAL 2.5W / 2.0W
$R_{DS(ON)}$	N P	250mΩ / 180mΩ 500mΩ / 350mΩ	130mΩ 200mΩ	70mΩ 80mΩ	SINGLE / DUAL 30mΩ / 70mΩ 50mΩ / 100mΩ
BV	N P	20V 20V	20V 20V	20V 20V	SINGLE / DUAL 20V 20V
I_D	N P	1A / 1.2A 0.7A / 0.85A	3.2A 2.7A	4.7A 4.2A	SINGLE / DUAL 7.9A / 4.6A 6.5A / 3.8A
TARGET MARKETS		Notebook Cellular Phone PDA Pager	Notebook Cellular Phone PDA PCMCIA	Notebook Cellular Phone HDD PDA	Notebook Cellular Phone HDD PDA
TECHNOLOGY	5M cell/in²	2.7V Gate Drive	2.7V Gate Drive	2.7V Gate Drive	2.7V Gate Drive

VIDEO SIGNAL TRANSMISSION AND DISPLAY

Video data has become a very important component of communications in the modern world. Applications range from entertainment, surveillance, heads-up displays, personal computing to high speed data links, to name but a few. In this section we will look at devices designed to handle video signals, both digital and analogue video signals, from transmission and routing through to eventual display.

Video is often digitised to enable signal compression for maximum use of cable or other wideband links. The Comlinear product group of National Semiconductor has introduced a number of products designed to cope with the losses and signal degeneration of digital video signal introduced by long lengths of commonly used cables.

Dealing with Lossy Cable Transmissions?

Lossy Media:

- *Coaxial Cable*
- *Unshielded Twisted Pair*
- *Or other media with similar dispersive loss characteristics*

2

Signal loss or attenuation is associated with any length of cable. Coaxial cable and twisted pair are examples of lossy media.

This cable attenuation does not allow high frequency transmission without compensation. The general equation for Belden 8221 cable loss is:

$$A = K \cdot L \cdot \sqrt{f}$$

where

$$K = 8.8 \times 10^{-6} \frac{dB}{m \cdot \sqrt{Hz}}$$

(for 75W Belden 8221 cable)

A = attenuation in dB

f = data rate in bits per second

L = length of cable in meters

Apart from the signal loss, the relatively low characteristic impedance of the cable presents a challenge at the driving end, particularly at the high data rates that may be employed.

National's Serial Digital Cable Products are the Answer

- *CLC006/CLC007 - Cable Drivers*
- *CLC014 - Adaptive Cable Equalizer*
- *CLC016 - High Speed Data Retiming PLL*
- *CLC018 - 8 x 8 Crosspoint Switch (Q4 1996)*

3

With this group of products, complete serial digital video (SDV) routers can be built using cable sections as long as 300m. The CLC006/CLC007 are able to provide full swing into matched cable loads, the CLC014 automatically compensates for cable losses, and the CLC016 is used to retime data and clock signals after transmission through the cable. Where multiple data sources are used, the soon to to be available CLC018 cross-point switch facilitates signal routing.

CLC006/007
Serial Digital Cable Drivers

CLC006 - 2 amplitude adjustable outputs

CLC007 4 outputs (2 pairs)

- *No external pull-down resistors*
- *650ps rise and fall times*
- *Operates from a single +5 or -5.2 V supply*
- *Low Power Dissipation*
- *DC to >400Mbps*

4

The CLC006 and CLC007 are the monolithic cable drivers. These ECL logic level drivers conform to the SMPTE 259M standard for the transmission of serial digital video signals. These devices can drive up to 300m of Belden 1505 or 8281 cable and deliver full ECL swings to a matched 75ohm load up to 400Mbps.

The CLC006 and CLC007 output stages and do not require pull-down resistors. This feature allows for lower power consumption and cost.

Potential Applications:

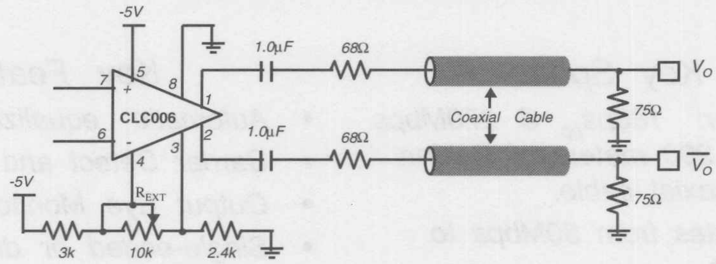
Digital video routers

Cable driver for digital data transmission

CLC006 - Low power replacement for GS9008 in most applications

CLC007 - Low power replacement for GS9007 and MC10EL89

Cable Driver with Adjustable Drive Level



R_{EXT}	V_O (V_{p-p})
10k	1.07V
5k	950mV
2k	760mV
1k	615mV
0	370mV

Both the CLC006 and CLC007 are able to drive a matched cable with an 800mV(p-p) signal without external components. The CLC007 has four outputs (two pairs), while the CLC 006 has two outputs. Since both devices are available in the same 8 pin package, the ‘spare’ pins on the CLC006 offer an opportunity to adjust the output drive amplitude. By connecting a resistor between pins 3 &4 , the drive level can be reduced from 800mV (open circuit) to 360mV (short circuit). If two additional resistors are used, connected to ground and the negative supply voltage rail, the output amplitude can be changed from 1.07v to 370mV.

CLC014
Adaptive Cable Equalizer

Key Specs

- *Low jitter: 180ps$_{pp}$ @ 270Mbps through 200 meters of Belden 8281 coaxial cable.*
- *Data Rates from 50Mbps to 650Mbps*
- *Low supply current: 58mA*

Key Features

- *Automatic equalization*
- *Carrier Detect and Output Mute*
- *Output Eye Monitor*
- *Single-ended or differential input*
- *All NRZ Data*

The CLC014 adaptive cable equalizer is a low-cost monolithic solution for equalizing data transmitted over cable (or any media with similar dispersive loss characteristics). This equalizer automatically adapts to equalize any cable length from zero meters to lengths that attenuate the signal by 40dB at 200MHz. This corresponds to 300m of Belden 8281 or 120m of Category 5 UTP (unshielded twisted pair).

Input Interfacing: The CLC014 accepts differential or single-ended inputs.

Output Interfacing: The outputs DO and DO produce ECL logic levels when the recommended output termination networks are used. The outputs are taken off of the collectors of the transistors. Refer to the CLC014 data sheet for recommended interfaces for standard ECL families.

Potential Applications:

SMPTE 259M serial digital interfaces: NTSC/PAL

Serial digital video routing and distribution

Serial digital data equalization and reception

Data recovery equalization for ATM, Ethernet, CAD networks, medical, set top terminals, industrial video networks

CLC016
High Speed Data Retiming PLL

Key Specs

- Low jitter: $130ps_{pp}$ @ 270Mbps
- Data Rates from 40Mbps to 400Mbps
- Low supply current: 110mA
- ±5% VCO center frequency accuracy

Key Features

- Retimed Output
- Recovered clock output
- Automatic and manual rate select
- Carrier detect output
- Single-ended or differential input
- All NRZ Data

The CLC016 is a low-cost, monolithic, data retiming phased-locked loop (PLL). The CLC016 is designed for high speed serial clock and data recovery. The auto-rate select (ARS) feature of the CLC016 simplifies high-speed recovery in multi-rate systems. The ARS and the pre-configured external resistors for the CLC016 set 4 separate data rates which are automatically selected by the CLC016 to provide accurate, low jitter clock and data recovery. A single resistor sets the data range anywhere from 40 to 400Mbps.

Input Interfacing: The CLC016 accepts differential or single-ended inputs once an input voltage meets standard ECL logic levels.

Output Interfacing: The outputs DO and DO produce ECL logic levels when the recommended output termination networks are used. The outputs are taken off of the collectors of the transistors. Refer to the CLC016 data sheet for recommended interfaces for standard ECL families.

Possible Applications:

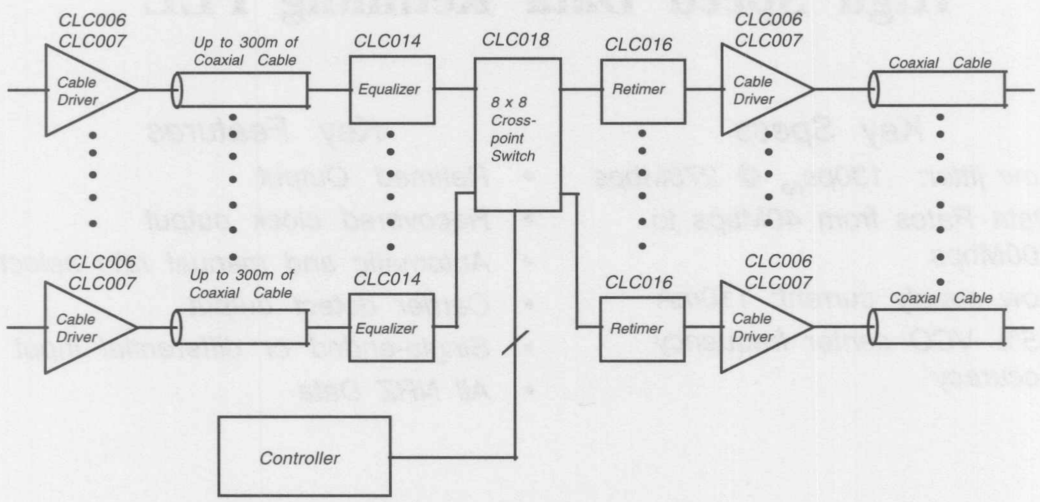
SMPTE 259M serial digital interfaces: NTSC/PAL

Serial digital video routing and distribution

Clock and data recovery for high-speed data transmission

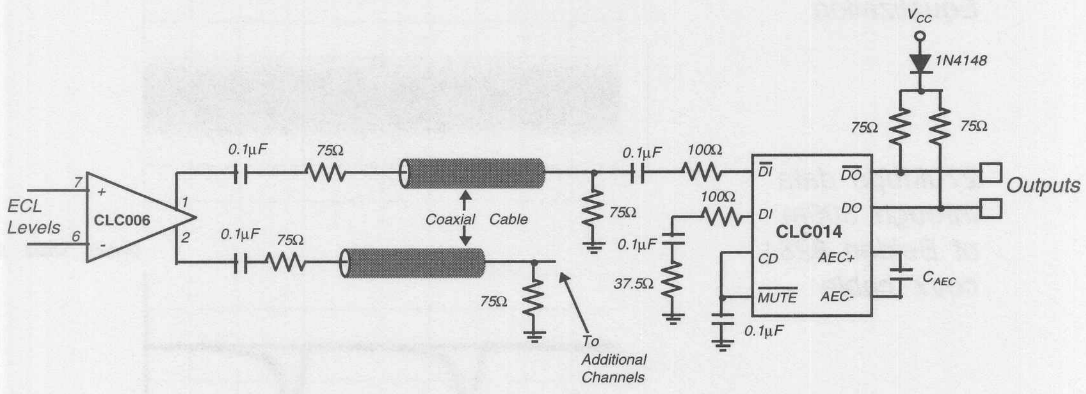
Re-synchronization of serial data for high-speed ATM, CAD networks, and medical or industrial imaging

Video Routing Block Diagram



This is a typical block diagram for a serial digital video (SDV) router. The CLC014 automatically equalizes cable lengths from zero meters to 300 meters at 360MHz. The equalized outputs are connected to the differential inputs of the CLC018 (available Q4 of 1996). The CLC016 receives the data from the crosspoint switch and performs clock and data recovery, reducing jitter. Finally, the retimed data is driven into more coaxial cable by the CLC006 or CLC007.

CLC006, CLC014
Coaxial Cable Receiver



This diagram illustrates a typical coaxial cable receiver. This circuit will equalize a variety of coaxial cables up to lengths that attenuate the signal by 40dB at 200MHz.

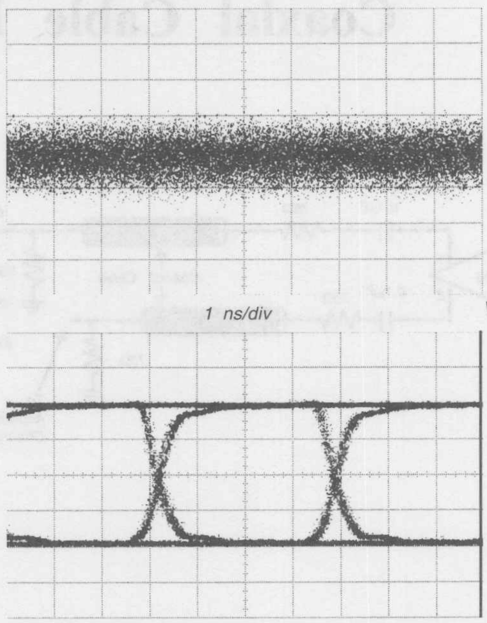
Circuit Description:

This illustrates the proper connections for a single cable driven by a CLC006 driver. Carrier Detect (CD) is connected to MUTE to latch outputs DO and DO in the absence of an input signal to the equalizer. This configuration prevents random output transitions due to noise.

Receiver Results

End of Cable before Equalization

270Mbps data through 300m of Belden 8281 coax cable



After Equalization

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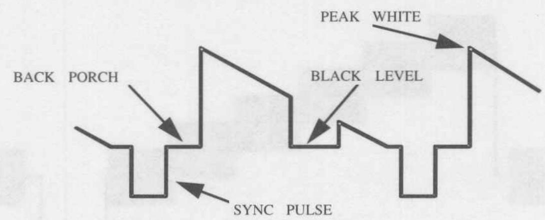
These plots show the performance of the CLC014 Adaptive Cable Equalizer. The top plot illustrates a 270Mbps data stream at the end of 300m of Belden 8281 cable, before equalization. The bottom plot shows the result or eye pattern of the CLC014, after equalization. The CLC014 will equalize data up to 650Mbps.

Analog Video Transmission

NTSC

PAL

SECAM



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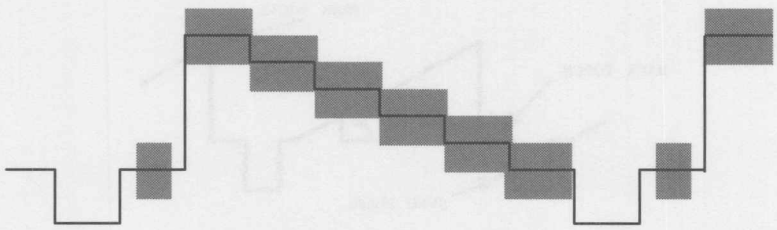
Up to this point we have been considering video transmissions where the video is in the form of a high bit-rate serial data stream. For many video applications the video waveform is in the more traditional "baseband" format, familiar to anyone that has worked in the commercial television environment. For a colour signal, the video is usually composed of three separate channels of information - the RED, GREEN and BLUE channels. Sometimes, as shown above, the sync signal is added to the GREEN video signal (sync on Green). In this particular instance we have "positive" video and "negative" sync relative to "black" level. Subsequent processing may change the RGB format to composite, luminance or colour difference components (Y, R-Y, B-Y, U/V, C, etc), but the demands on the signal processing devices used in the transmission channel do not change very much with format changes. In this next section we will briefly review some of the I/Cs suitable for analogue video processing.

Amplifier Requirements for Analog Video

- Wide Bandwidth
- High Slew Rate
- Low Differential Phase
- Low Differential Gain



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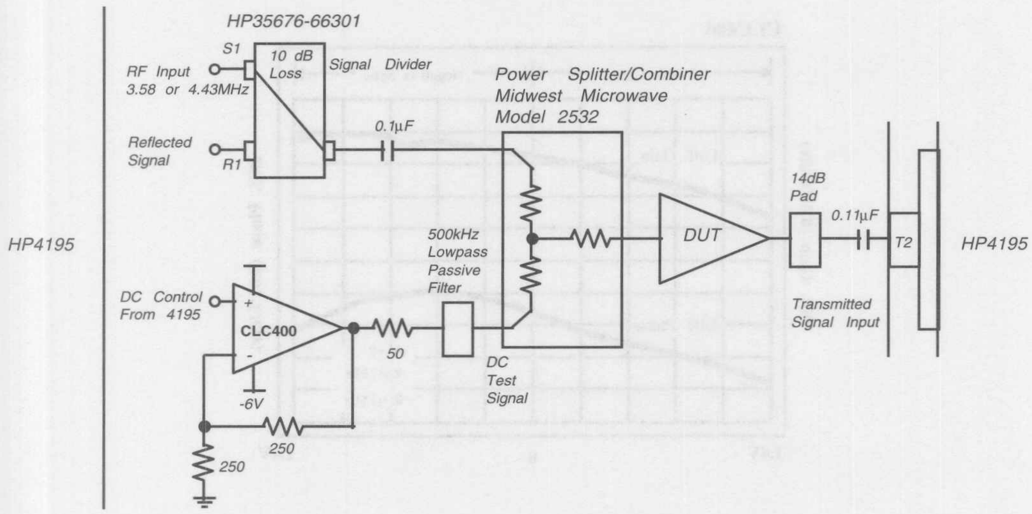
As might be expected, for good fidelity the closed loop bandwidth of the amplifier must be significantly higher than the highest signal frequency - at least 60 MHz to 80MHz for dealing with the most commonly encountered analogue video formats. Further, this bandwidth must be available for relatively large signal swings into low cable impedances. Although the nominal video level is 1V, it is often desireable for the amplifier to deliver at least 2V to compensate for cable termination losses.

Two unusual specifications for a video amplifier are DIFFERENTIAL PHASE and DIFFERENTIAL GAIN. These refer to the ability of the amplifier to linearly process a low level high frequency signal at any dc level within the anticipated large signal amplitude range. These specifications can be measured using a special waveform, shown above, where a subcarrier (at 3.58 MHz or 4.43MHz) is superimposed on a stepped video waveform that takes the subcarrier average level from white to black. Differences in the amplitude and phase of the subcarrier at each of the stepped levels is detected by a Vectorscope and the worst case difference becomes the specified performance level. Most vectorscopes are able to resolve <1 degree DP and <1% DG.

Using A Network Analyser for DP & DG Measurements



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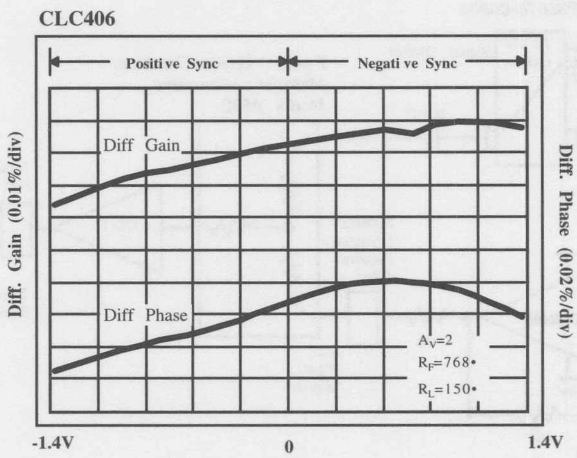
When we are measuring amplifiers for use in a video system, the amplifier's DP and DG have to be considerably better than the overall system specification, since many such amplifiers may be included in the signal path. If a network analyser, such as the HP 4195, is used to make the measurement, DP can be resolved to significantly less than 0.01 degrees and DG to less than 0.01%. Unlike the stepped technique shown previously, the subcarrier is ramped through the entire measurement range and the levels of DP and DG recorded at 21 points. Both positive and negative video polarities are accomodated with this measurement technique.

CLC406 DP/DG Performance



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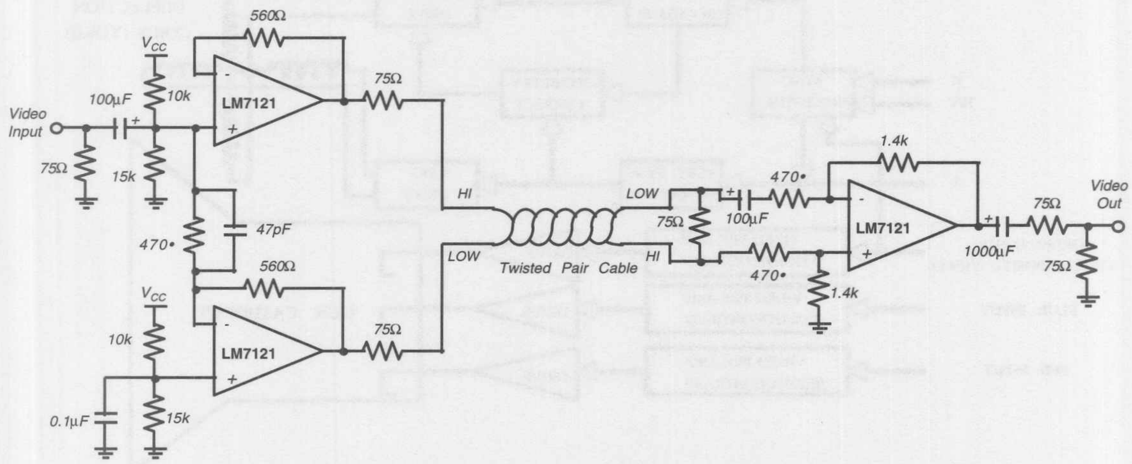
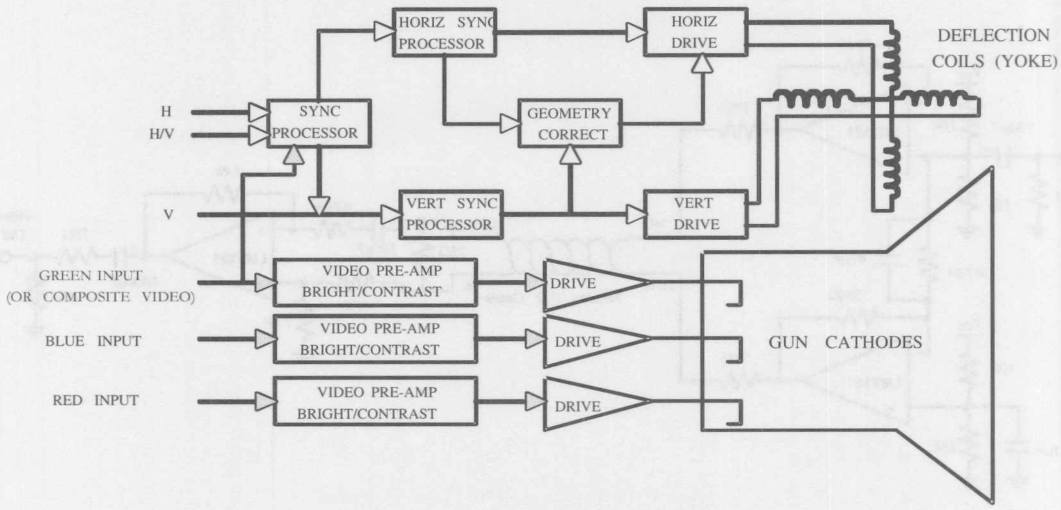
These curves clearly illustrate the advantages of this measurement technique. Apart from the higher resolution compared to a vectorscope, the curves show how the amplifier is performing throughout the entire video dynamic range. Also, it is obvious from the DP curve that simple end-point measurements would be misleading about the actual performance of the part.

Video on a Twisted Pair



Not all applications for video are as demanding. For example in this circuit, shown previously in the Amplifier section, a number of devices are not being cascaded. In fact the overall system requirements on DP and DG could be as high as 5 degrees DP and 10% DG. Here the typical performance of the LM7171 will yield less than 1 degree and 1%. More important is the ability of the LM7171 to drive the load represented by the twisted pair cable.

MONITOR SIGNAL PROCESSING



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Because they share the same display device, a CRT or cathode ray tube, many of the early monitor designs were similar to television receiver circuits, particularly with regard to the design of the deflection circuits. The rapid introduction and popularity of colour displays has changed this, with many new I/Cs being developed specifically for colour monitors with display resolutions as high as 1024 X 768 pixels. National has been a leader in the development of these circuits, and this next section will focus on the most recent I/Cs designed for monitor applications.

The block diagram above shows the major circuit elements of an RGB monitor; the video pre-amplifiers which raise the input signal level from about 1Vp-p to 4Vp-p, and allow user adjustment of screen brightness and contrast; the CRT drive amplifiers that amplify the signal to the 40Vp-p to 60Vp-p level required at the cathodes of the CRT; the sync processing circuits that identify the horizontal scan and vertical scan components which position the electron beams on the face of the CRT (ie generate the raster); and finally the geometry correction circuit that compensates for scanning errors produced by the deflection circuits and the CRT gun structure and "optics".

Before looking at any of these circuits in general, it is worth noting that while the three channels for processing the video (one each for the RED, BLUE and GREEN components of the video signal) are fairly common to most colour monitors, there are several ways of handling the synchronisation signal. This can be provided as separate horizontal and vertical sync H & V (with either positive or negative polarity), as composite sync where both horizontal and vertical sync are in the same waveform H/V, or as a composite sync signal added to the GREEN video signal, G + S.

VIDEO PREAMPLIFIERS

Single Channel Amplifiers:

LM1201	130 MHz
LM1202 (with Master/Slave Contrast)	230MHz

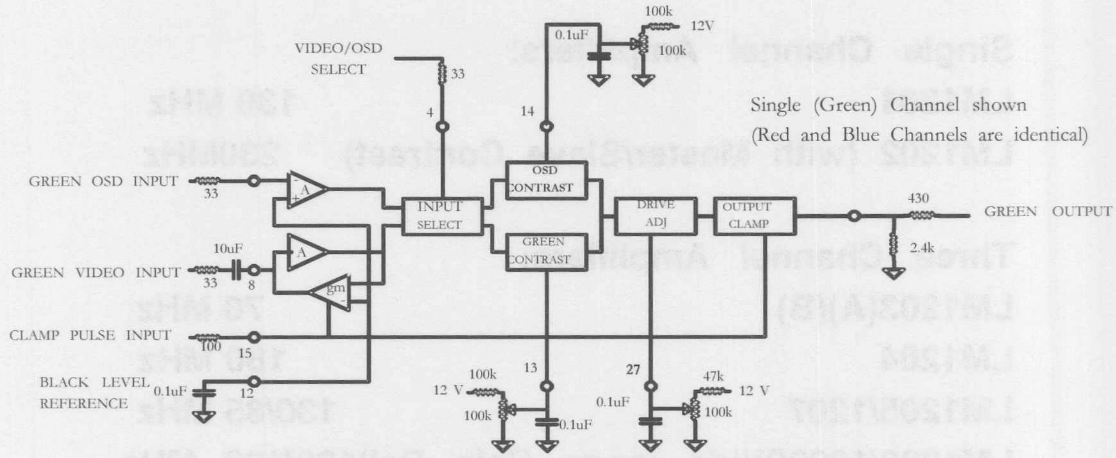
Three Channel Amplifiers:

LM1203(A)(B)	70 MHz
LM1204	180 MHz
LM1205/1207	130/85 MHz
LM1208/1209(Wide range Gain Bal)	130/100 MHz
LM1281 On Screen Display (OSD)	85 MHz

National manufactures a number of video pre-amplifiers capable of processing all of the most popular PC video display resolutions. For extremely high resolution (work station displays), or for monochrome displays, there are two devices, the LM1201 and LM1202. The LM1202 allows the contrast section to be slaved to other devices so that tracking is obtained when three separate I/Cs are used for RGB operation

For most RGB applications, the RED, GREEN and BLUE channels are combined into a single I/C for best matching between channels and to ensure that the colour balance is maintained through the full range of user adjustments (ie contrast and brightness settings). The LM1203 was the first product in this category, which has been expanded to include the LM1204 and the LM1205 series. The latest product is the LM1281, which adds a 70MHz OSD (On Screen Display) capability. The TTL compatible OSD inputs allow visual confirmation of display settings and changes in levels by the user.

LM1281 Pre-Amp with OSD



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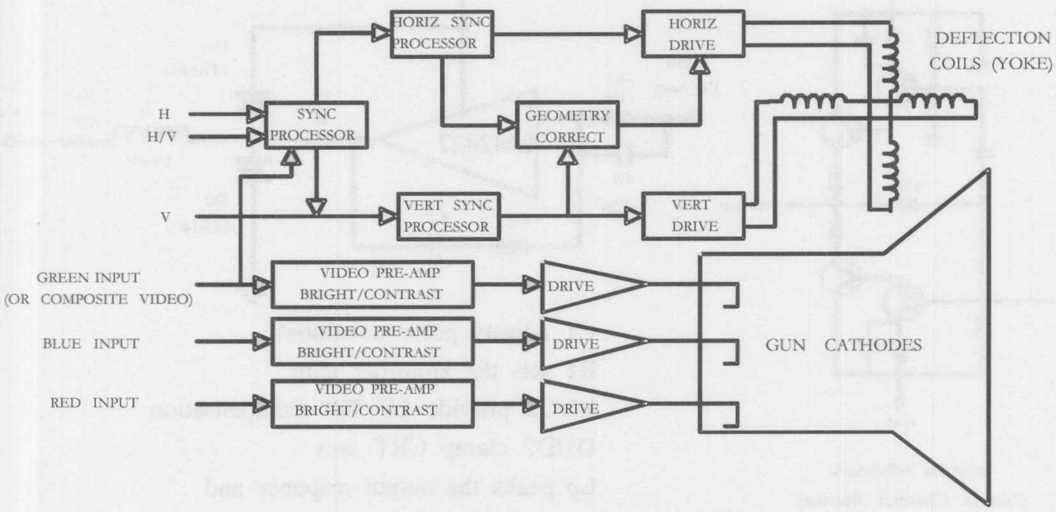
The video input signals is ac coupled into the video pre-amplifier so that most types of video sources can be accomodated. However for proper display of the video brightness level, the signal has to be dc restored and, in the case of a colour display, matched to the proper level compared to the other two channels for the correct screen colour balance to be obtained.

Shown here is a single channel of the LM1281. The BLUE and GREEN channels are identical and share common control of the Contrast (signal ac amplitude), Pins 13 & 14. The OSD input is dc coupled and buffered to the input select switch that determines whether the video or the OSD will be displayed on the screen. The buffer amplifier also sets the input black reference level since internal feedback will ensure that the non-inverting input will track the OSD signal level. When the OSD input is low or at black level (during the display blanking period), this level is applied to the gm stage in the feedback loop of the green video input amplifier. Gating the gm stage on during the blanking period will store this dc level on the capacitor at Pin 12. The ac coupled green video signal will be dc restored to this level, and if the clamp pulse timing co-incides with the back porch period of the video, the black level for the OSD input and the video will be matched for each channel. Contrast controls at pins 13 & 14 set the OSD and video contrast levels respectively for all three channels with a 40dB range for a 0-4Vdc control voltage. This dc voltage can be supplied from a potentiometer as shown, or from the output of a 5V DAC.

Channel gain matching over a 12dB range is provided by similar dc voltages at Pin 27 (26 & 28 for the Blue and Red channels). The output stage is also clamped with separate dc level storage capacitors so that the CRT white level or brightness can be controlled.

If necessary, the output stage can have a blanking pulse applied (not shown), which drives the output down to less than 500mV, which will ensure CRT gun cut-off during the retrace part of the signal. If the blanking input is not used it is held high with a resistor to the supply.

CRT DRIVERS



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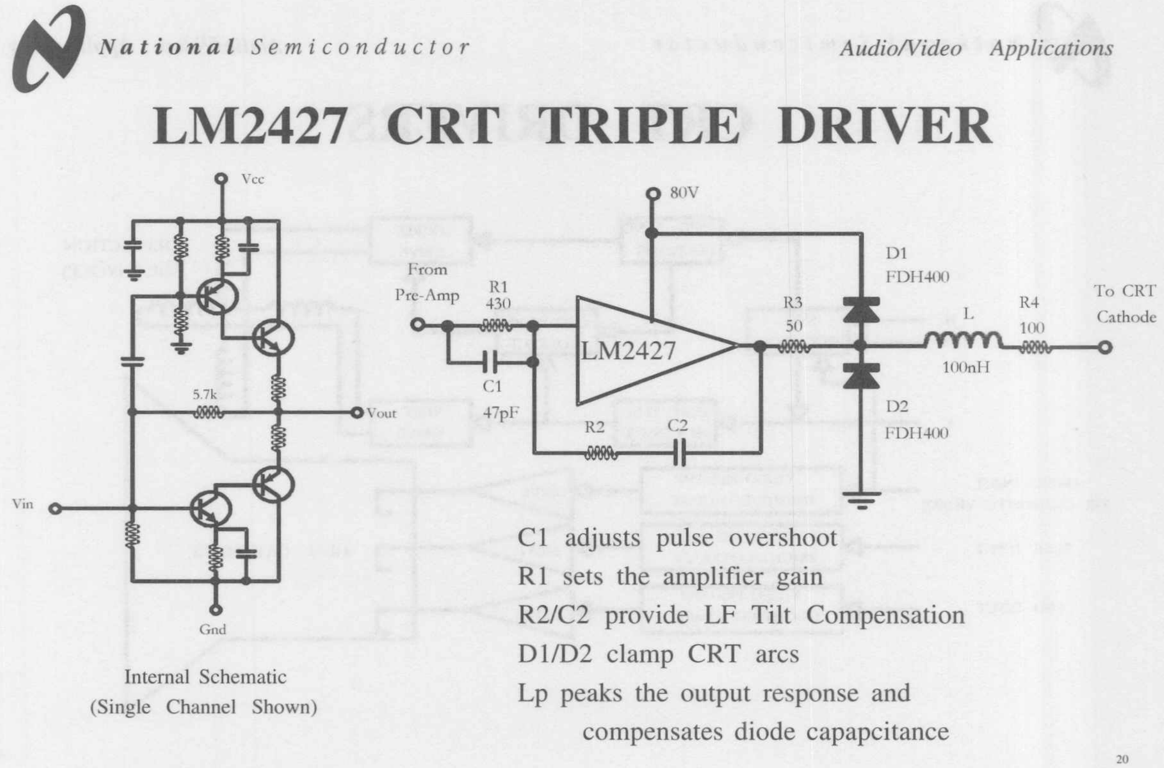
Once the input video signals have been matched, the next step is to raise the amplitudes to the level required to directly drive the CRT cathodes. This typically means having the capability to drive a capacitive load (from 8 to 15pF), at amplitudes up to 60Vp-p, over a frequency range from 30MHz to over 200MHz. Traditionally this area has been dominated by discrete designs, with concomitant problems of channel matching, physical size, and emi radiation. Under 100MHz this situation has been alleviated by Hybrid designs, which put all three channels on a single substrate. The smaller physical size of a single package, easier heatsinking and radiation shielding, has made the production of neck-mounted video driver boards easier. Now National is taking the process a step further with the introduction of monolithic CRT drivers.



First, however, lets look at a typical Hybrid CRT video driver, to see how they work and what some of the constraints are.

The LM2427 is a Triple video driver with all three channels in a staggered lead TO-220 style package. The Tab of this low cost plastic package makes heatsinking (a requirement) fairly straight forward and only a single heatsink is required for all three channels.

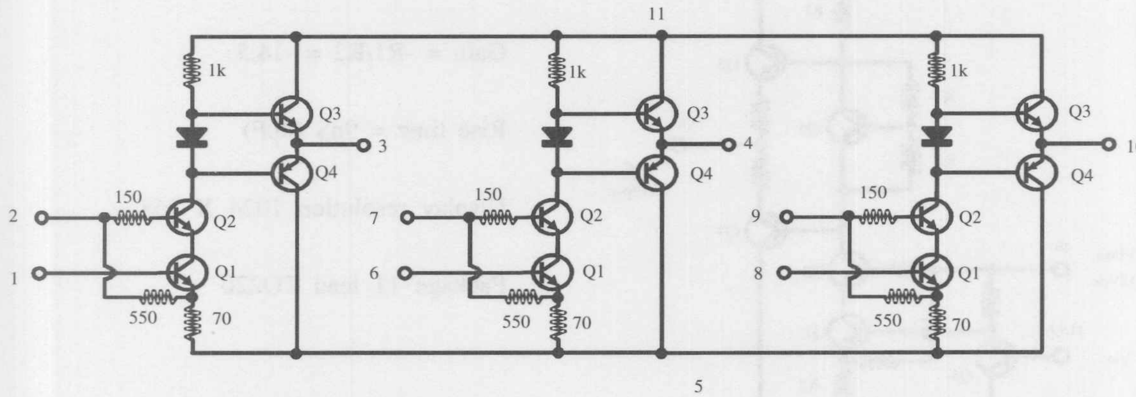
An internal 5.7Kohm resistor is fed back to the input from the push-pull output stage. This resistor, along with the video source resistance, will set the inverting gain for the amplifier. For example, if the resistor in series with the video pre-amp output (assumed to be low impedance buffer), is 430 ohms, the inverting gain is $-5.7X10^3/430 = -13$. With a pre-amp output level of 4Vp-p, this is enough gain to provide 50Vp-p at the cathode.

Although many CRT drivers are specified in terms of bandwidth capability (eg 80MHz at 50Vp-p), for the driver the pulse rise and fall times are also considered to be important. The LM2427 is able to deliver rise and fall times in the vicinity of 3.5nS. Nevertheless, achieving these rise and fall times is not totally dependent on the device itself. As the capacitive load is increased so will the rise time. For example at 50Vp-p the rise time increases to 4.5nS with a 13pF load (which can be contributed to by lead lengths and protection components as well as the CRT gun structure). A small capacitor across the input resistor can be employed to produce signal overshoot, thus improving the rise time. Although component values are suggested in the data sheet, the proper value is often found by experimentation.

Sometimes the output waveform will exhibit “tilt”, which will show on the screen as a smear following a large area of black. This is because large areas of uniform brightness are low frequency signals causing high power dissipation which is not uniformly shared in the amplifier’s transistors. The R2/C2 network is designed to provide compensation for tilt.

The clamp diodes and output resistors are intended to prevent damage to the CRT drivers from arcs (the CRT will also have spark gaps for this). Since these components will add to the output capacitance, a small inductor is used to provide peaking. Again the value is usually determined by experiment with a specific circuit.

LM2419 65MHz TRIPLE DRIVER



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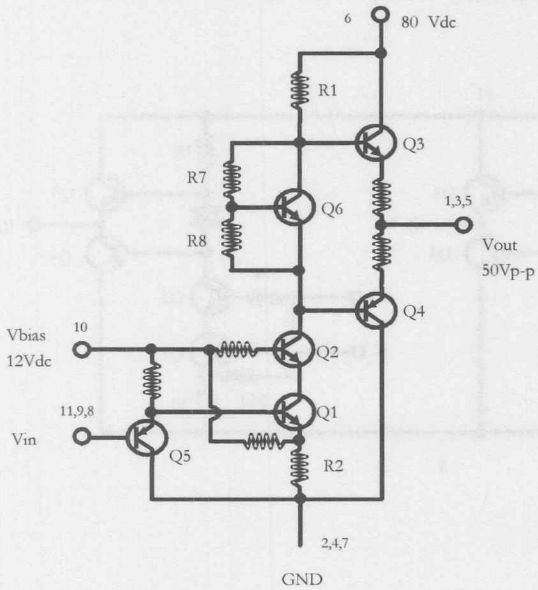
The LM2419 is another triple CRT driver in the 11 lead TO-220 package. This device has a slightly different design structure, and the overall signal gain is set internally. The input stage Q1,Q2 provides all the gain and the output Q3,Q4 is a unity gain push-pull stage. At the input Q1 gain is set by the ratio of the collector and emitter resistors to -15. Q2 is a cascode stage with the base heavily bypassed to a DC voltage (12Vdc), to help isolate Q1 collector from the capacitive load represented by the output stage. A diode between the push-pull stage bases helps to minimise the crossover distortion as conduction switches between Q3 and Q4.

For this and the previous amplifier, power dissipation can be significant, and is highest when a white screen is displayed (amplifier output low). With about 20% of the time spent in retrace, the average power dissipation at 50Vp-p will be close to 4Watts/channel. If the device case temperature is limited to 90°C and an operating ambient of 50°C is anticipated, then the thermal resistance of the heat sink must be at least 3°C/Watt ({90-50}/12).

LM2406 Monolithic TRIPLE



Gain = -R1/R2 = -14.5

Rise time = 9nS (8pF)

Display resolution 1024 X 768

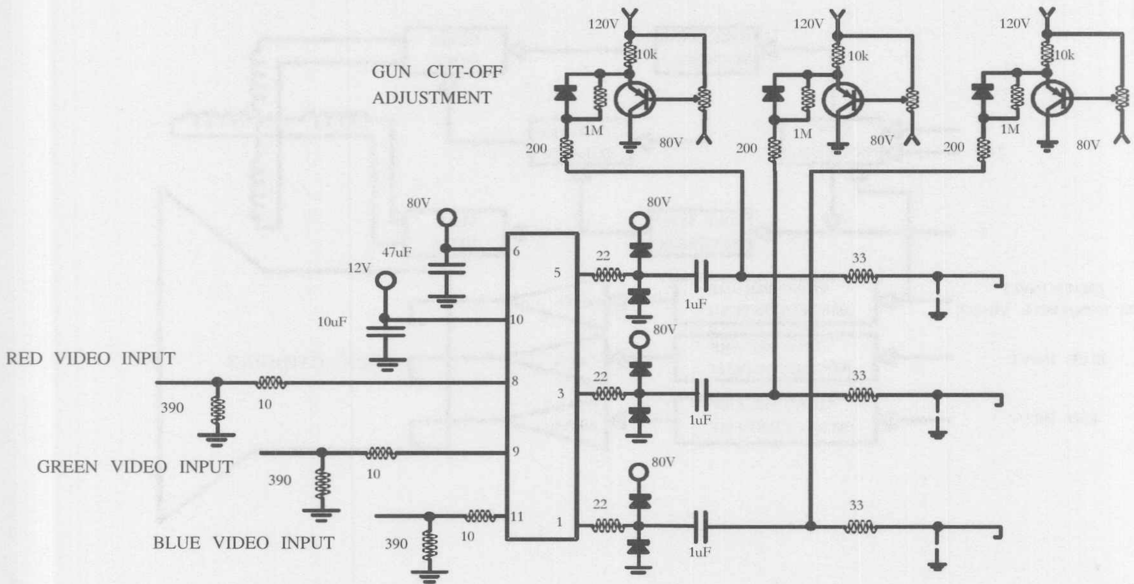
Package 11 lead TO220

22

The latest addition to National's portfolio is the LM2406 MONOLITHIC triple CRT driver. Just as the previous Hybrid devices, the LM2406 is able to deliver 50Vp-p operating from an 80V supply. The rise and fall times are typically 9nS with an 8pF load, making the driver suitable for VGA and SVGA displays.

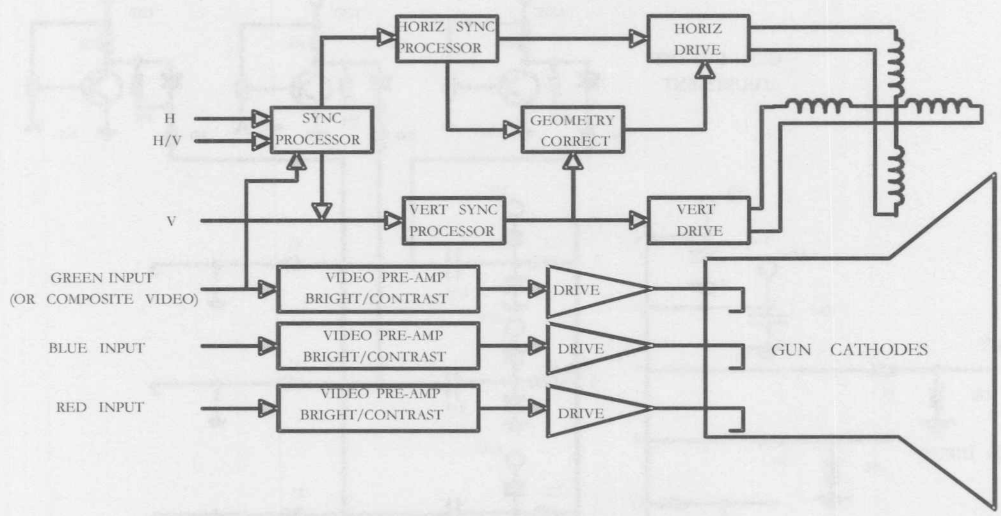
As the internal schematic shows (one channel only shown), the topology is similar to the LM2419. An input cascode stage is used, Q1 & Q2, to provide an internally set gain of -14.5 and to isolate the load capacitance. Instead of a simple diode, the crossover bias is provided by Q6 which is set up as a Vbe mutiplier to exactly match the combined base-to-emitter voltage drops of the output NPN & PNP transistors Q3 and Q4. Momentary short circuit protection (<1 sec) has also been provided in the output stage. The power dissipation under peak white conditions is slightly less, 3.5 Watts/channel, allowing smaller heatsinks to be used, around 4.75°C/W in an operating ambient of 50°C.

TYPICAL APPLICATION LM2406



In this application the pre-amp supply voltage (12Vdc) is used as the bias point for the base of all three cascode stages and is bypassed close to the package. Each output stage has diode arc protection in addition to the spark gaps on the CRT cathodes. Notice that although the video signals are dc coupled from the pre-amplifier stages (LM1205 outputs for example), the driver output is ac coupled to the cathodes. This is done to allow the cathodes to be biased to a higher dc voltage (120V) than the supply voltage for the driver (80V). A higher voltage range allows a wider gun dc cutoff voltage variation yet still allows the full ac signal voltage swing. On the cathode side of the output capacitor, the gun dc voltage is re-established through the clamp diodes and transistor voltage sources.

SYNCHRONISATION



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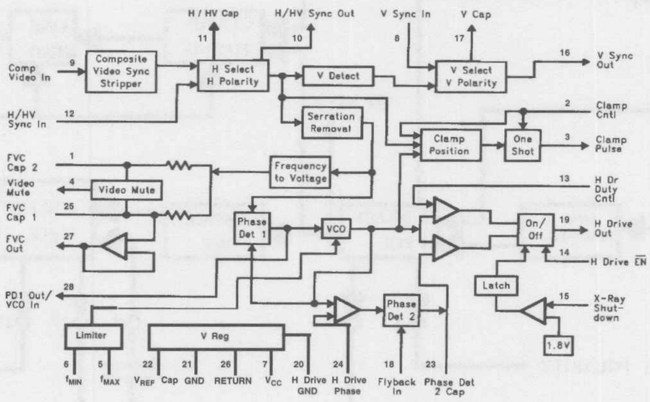
Another key monitor function is to synchronise the deflection circuits (which control the position of the electron beams) with the video source. As noted earlier, the source sychronising signal can come in several different ways. Totally separate horizontal and vertical sync signals may be provided. Usually these are TTL level and negative polarity, but sometimes positive polarity will be encountered. Similarly, a combined horizontal and vertical sync signal will usually be negative polarity but may also be reversed. For some monitors the sync signal will be added to one of the RGB channels, usually GREEN, where the amplitude will be substantially less than TTL levels. The LM1292 has been designed to accomodate all these variations, and has an on-board oscillator to provide a default sync signal in the absence of external video or sync. This is important for on screen displays for diagnostic purposes and for generating the HV for the picture tube anode.

LM1292 SYNC AND PLL



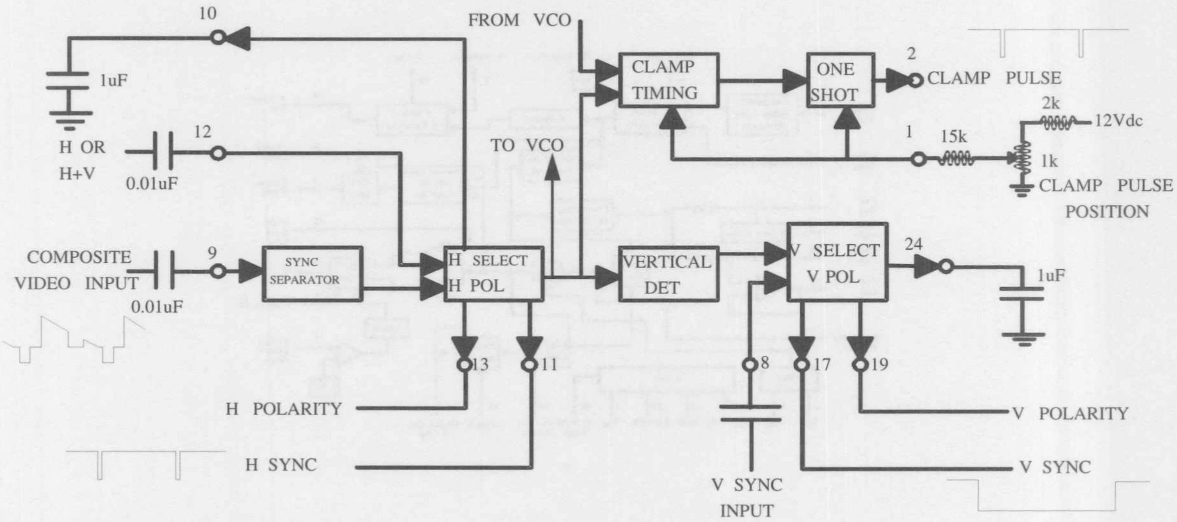
25

The block diagram for the LM1292 shows the major elements and Pin assignments for this I/C. It can be further divided up into two main sections; the sync processing section and the horizontal PLL section.

The sync section detects the type and polarity of the sync information that is being supplied, separates the horizontal and vertical sync components, provides flags indicating the sync polarity and helps to generate the appropriate clamp pulses for the monitor video amplifiers described previously.

The horizontal PLL includes a wide range VCO that can be locked to scan frequencies from 22KHz to as high as 125KHz, covering the all the popular scanning standards. The output driver stage has an adjustable pulse width which can be phased independently of the VCO output phase. This allows the video to be centred on the CRT raster and any necessary geometry corrections made to produce straight vertical edges to the picture. The horizontal output can be disabled by a TTL level signal (power conservation for example) or by an input from the HV section. If the HV rises to dangerous levels that could cause X radiation, a voltage divider from the HV will cause the horizontal output to shut down. Since the horizontal sweep is used to generate the HV, shutdown will remove the dangerously high voltage. A latch prevents the drive from starting up again until the problem has been resolved.

LM1292 SYNCHRONISATION IDENTIFICATION AND SEPARATION



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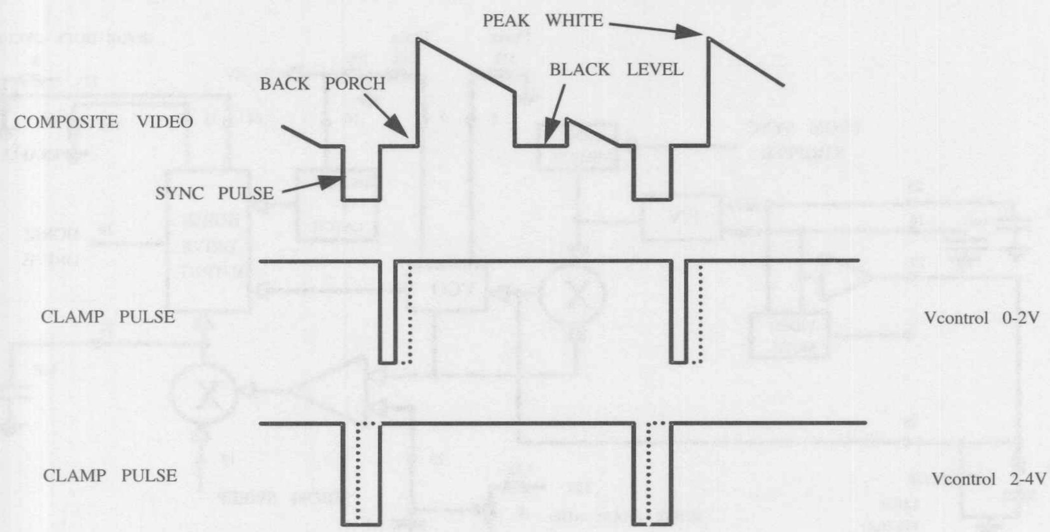
Here is the sync section in more detail. If horizontal and vertical sync of more than 1Vp-p amplitude are available these have priority over composite sync or composite video inputs. Two flags at Pins 13 & 19 will give the polarity of the incoming sync waveform from any source, a high level at either pin indicating negative polarity polarity sync. Regardlesss of the incoming sync polarity , the outputs at Pins 11 & 17 will be TTL and CMOS compatible negative polarity waveforms.

If the input source is composite video, the signal is again ac coupled to Pin 9. An internal clamp will restore the negative polarity sync tips to 2Vdc and then strip out the sync portion of the waveform. An incoming sync amplitude of at least 150mV is required if video contamination of the sync is to be avoided. This corresponds to an input video amplitude greater than 0.5Vp-p.

The horizontal sync is used both to synchronise the internal horizontal oscillator and to generate a video clamp pulse. Since both the timing (ie the start) and the width of the clamp pulse will depend on system requirements, the LM1292 is able to adjust these two parameters based on a control voltage at Pin1.

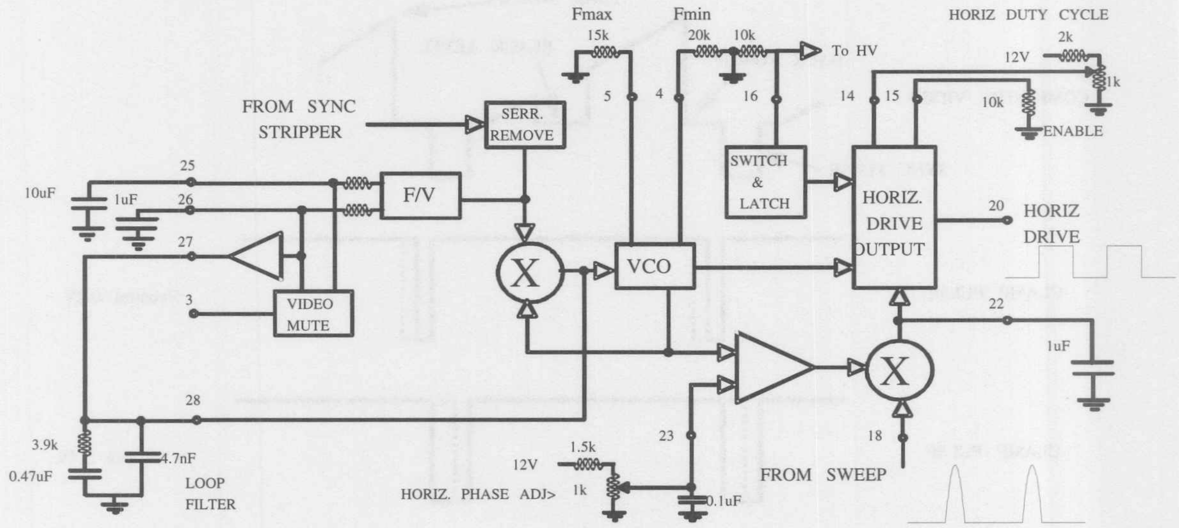
LM1292 CLAMP PULSE TIMING



27

The sync output and the VCO waveform are compared in the clamp circuit with the voltage at Pin 1, setting the timing position of the clamp pulse and the width of the clamp pulse generated by the one-shot. When the control voltage is between 0v and 2V The clamp pulse will start just after the trailing edge (positive going edge) of the incoming horizontal sync signal. As the control voltage increases the width of the clamp pulse will increase. Since the clamp pulse occurs during the back porch period of the video blanking period, it can be used with the video pre-amp to accurately dc restore the video black level reference. The control voltage is also used to control the clamp pulse width since the back porch period varies with the different scanning standards that are in use. This prevents inadvertent clamping at the start of the active video. When the control voltage is greater than 2V the position of the clamp pulse moves to the start of the sync pulse (negative going edge) and further increase in voltage will decrease the duration of the clamp pulse. By using the VCO input as part of the clamp pulse circuit, a pulse will be generated even in the absence of a sync source, allowing the on screen display to still function properly in the absence of external video.

LM1292 PLL AND H DRIVE



28

The other use for the horizontal sync (actually the main use) is to lock the LM1292 VCO to the input frequency. This input frequency can be between 32KHz and 120KHz and the oscillator must lock to this range and produce a low jitter drive to the horizontal drive circuit.

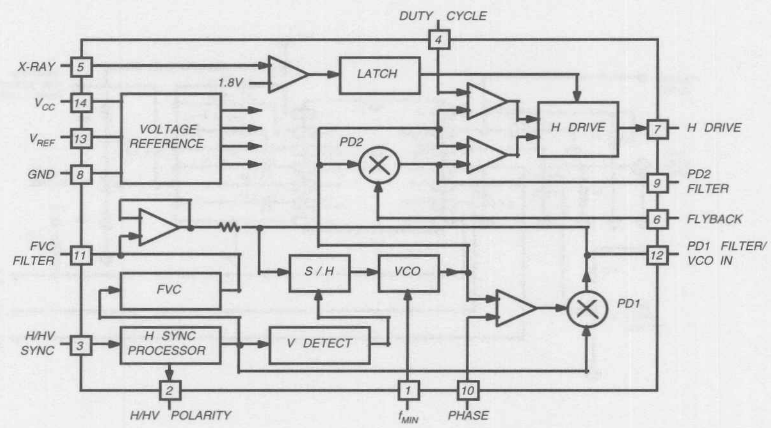
If the source is composite video or composite sync, the serrations are removed and the resulting waveform applied to one input of a phase detector and to an F/V circuit. The F/V generates a voltage based on the incoming sync frequency and superimposes this on the VCO control voltage to put the oscillator into the right frequency range that the main control loop can cause the oscillator to lock. Two external resistors set the available oscillator frequency range for fixed or limited standard compatibility monitors.

The other input to the phase detector is the VCO output, with the phase detector output being filtered at Pin 28 to produce the proper loop response. This filter, along with the VCO/Phase detector loop gain will determine the lock-in range and time, and the response of the loop to noise or other perturbations. Once locked the VCO output jitter is less than 1.3nS at 100KHz.

If there is an abrupt change in the incoming source frequency (a switch to a different scan standard) the F/V will cause Pin 3 to go low (an open collector output) which change can be used to mute the video during the transition.

Once the oscillator has locked to the incoming frequency the output is used to generate a horizontal drive waveform. The output is compared in a second phase detector with a pulse feedback from the horizontal sweep transformer. The threshold level at which the phase comparison occurs is set by an external voltage at Pin23. This allows the video to be centred on the raster and, if necessary, adjustment of the video position during vertical scan. The output pulse width of the horizontal can be adjusted and the output shutdown for power conservation or safety reasons.

LM1290 AUTO SYNC I/C



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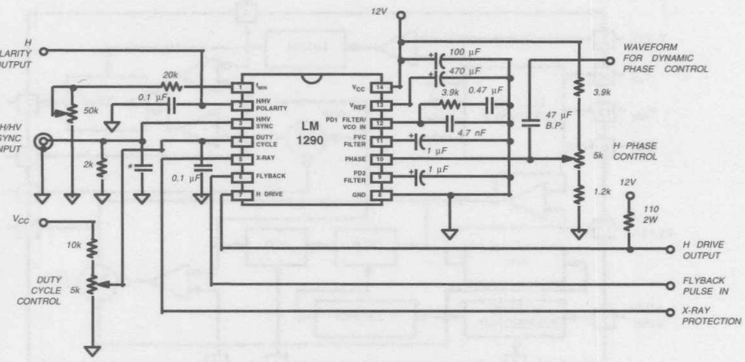
The LM1290 Horizontal deflection processor has many of the features of the LM1292, but is available in a 14 pin package. This device is designed to work with horizontal or composite sync (H or HV) and the VCO will automatically lock to any incoming sync frequency between 22 kHz and 90 kHz. An internally coupled F/V converter provides the dc shift to the main loop filter to change the centre frequency of the oscillator, and the rate of frequency change is contolled by a capacitor at Pin 11. This is used to prevent the on-time of the the horizontal output transistor from increasing excessively during a transition from a high to low incoming sync frequency. The drive to the horizontal output transistor can also be shut down in response to an excessive crt anode voltage. When the voltage at pin 5 (derived from a tap on the focus bleeder divider) goes above 1.9 Volts, the drive output is latched open until the supply voltage falls below 2V, ie the device has been turned off.

LM1290 APPLICATION CIRCUIT



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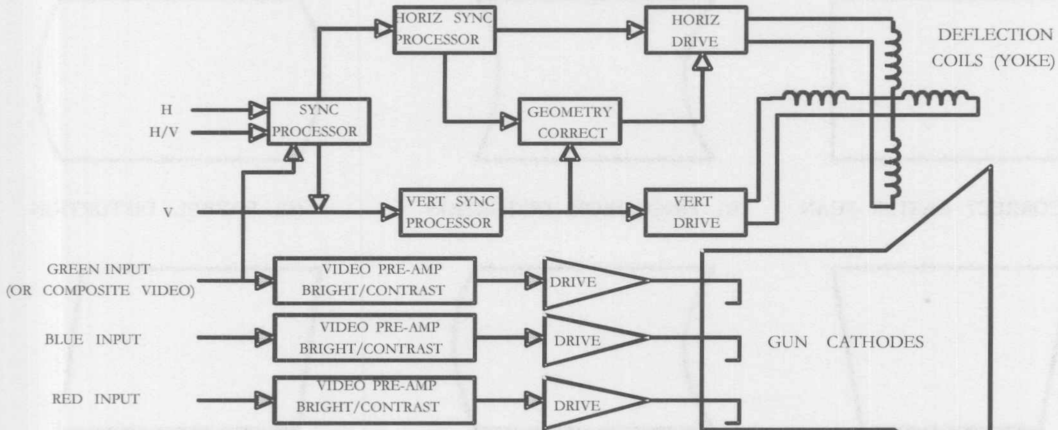
This circuit shows most of the external components required for a practical application of the LM1290. To maintain low jitter performance, the sync input components should be located as close as possible to Pin 3. The main loop filter, Pin 12, will determine the locking characteristics of the VCO, and its immunity to extraneous noise. Reducing the resistor size will damp the loop, producing slower lock times, but reducing the effects of noise. Increasing the resistor size will produce wider lock ranges and faster acquisition times. The oscillator phase in lock is controlled by the feedback pulse from the horizontal sweep transformer to Pin 6. However the actual drive waveform phasing can be changed with a DC control voltage at Pin 10 of LM1290. A dc voltage will affect the horizontal position of the screen display (Horizontal centering) while an ac voltage derived from the vertical deflection circuits will allow dynamic phase correction (covered later).

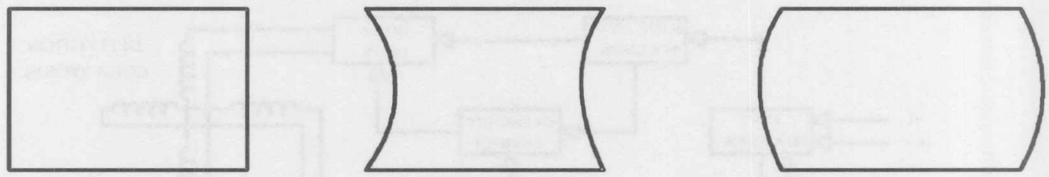
LM1295 GEOMETRY CORRECTION



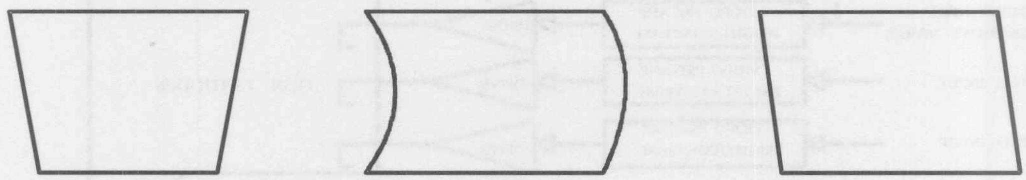
31

The final part of the monitor circuits to be covered is the vertical sync section. The LM1295 is a companion I/C to the LM 1292 and is used to provide a vertical scan signal in the absence of an external signal. It also provides vertical rate correction signals that are applied to the horizontal section to compensate for geometric distortion in the displayed picture. Since most monitor displays show the picture edges (unlike television receivers which are typically overscanned) any distortion is immediately apparent.

RASTER SCAN DISTORTION



(A) CORRECT RASTER SCAN (B) PINCUSHION DISTORTION (C) BARREL DISTORTION



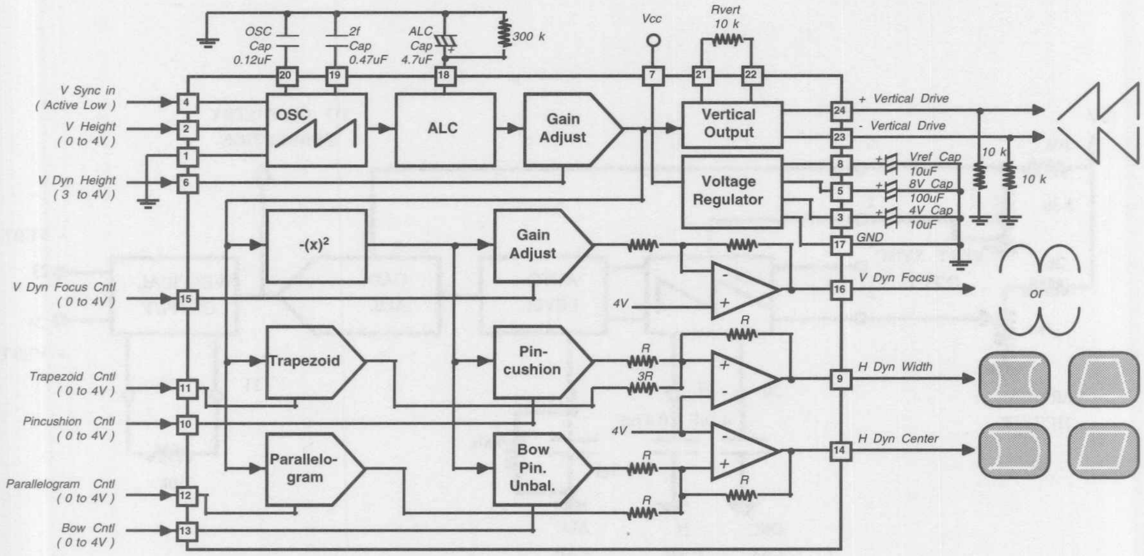
(D) TRAPEZOID OR KEYSTONE (E) HORIZONTAL BOWING (F) HORIZONTAL PARALLELOGRAM

32

Ideally the raster should be a perfect rectangle on the face of the CRT, centred from left to right and with straight edges. Unfortunately applying constant scan coil currents to the deflection coils will not always produce this result. One common problem is that as the beam moves off the centre axis of the tube the deflection angle increases and the electrons have farther to travel. If the horizontal deflection current does not change the scan lines will be shorter at the centre than at the upper and lower portions of the screen. This is known as Pincushion Distortion (and over correction is known as Barrel distortion) from its appearance on the screen. Sometimes the deflection coils can be imbalanced or improperly mounted on the neck of the picture tube. This can give rise to Keystone or Trapezoidal distortion where lines at the top of the screen are longer than those at the bottom. Most of the time the distortion is the result of several sources such that several correction factors have to be applied simultaneously. For example simple E-W correction for Pincushion can result in picture Bowing or tilted edges to the sides. Since these are (relatively) linear distortions, they can be compensated for by adjusting the phase of the Horizontal output during the vertical scan.

The LM1295 provides the vertical timing and wave forms necessary for these corrections to the horizontal scan.

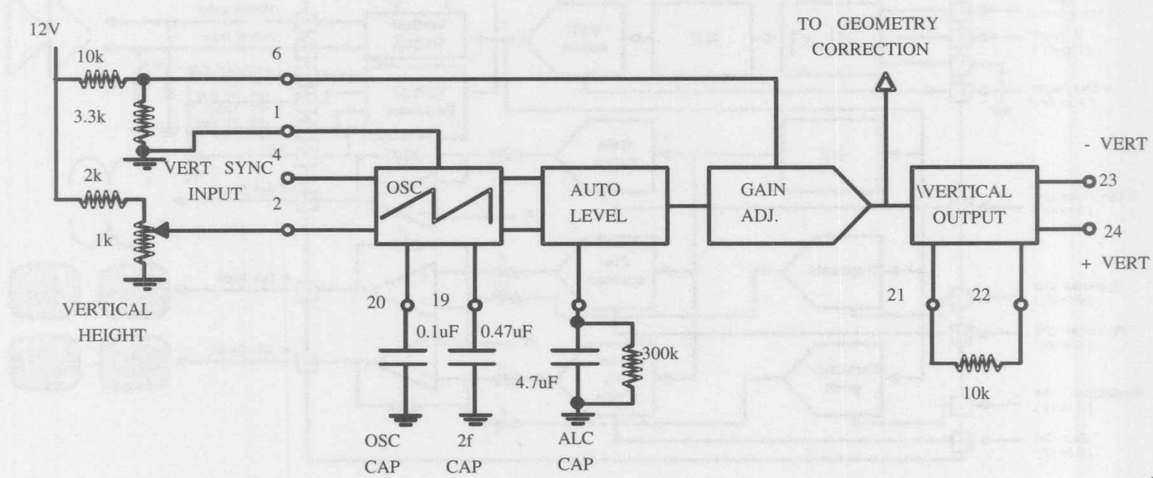
LM1295 APPLICATION CIRCUIT



33

From this block diagram we can see that the LM1295 can also be divided into two main sections. One section is the vertical oscillator and output driver circuit, and the other section a group of function generators and multipliers that are used to provide the correction voltages which are added to the horizontal scan for geometry correction.

LM1295 VERTICAL OSCILLATOR



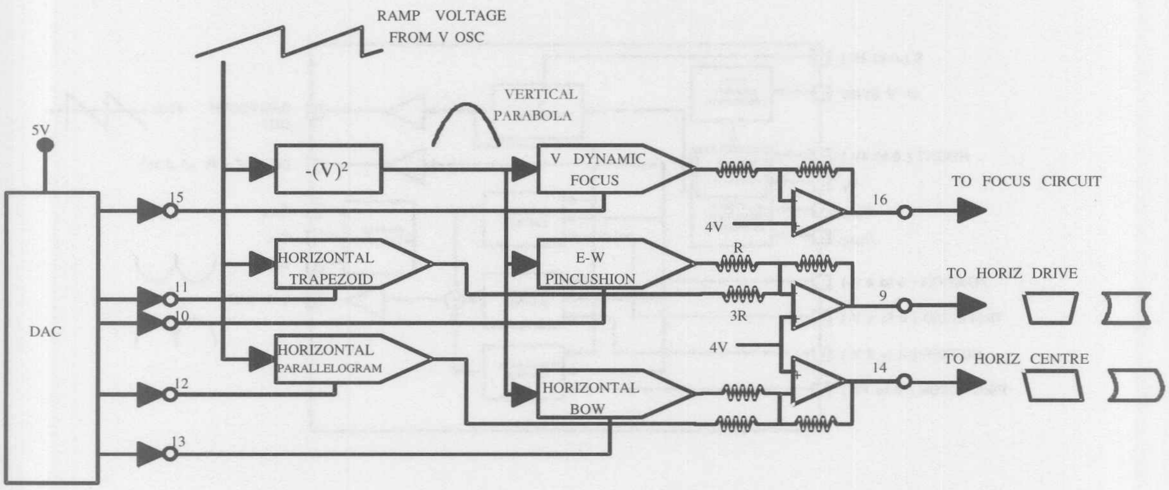
34

The vertical oscillator is an injection locked ramp generator with automatic level control. In the absence of TTL negative polarity sync, the oscillator will free run at 48Hz. This frequency is set to be just below the lowest anticipated vertical scan frequency by the 0.1uF capacitor at Pin 20. Incoming sync will injection lock the oscillator at frequencies up to 170Hz. A capacitor of half this value (in this case 0.47uF) at Pin 19 will prevent the oscillator from locking at twice the incoming frequency. The amplitude of the ramp can be controlled over a 1.8 to 1 range with a 4V to 0V control voltage on Pin 12. This voltage can be obtained from a potentiometer as shown, or from the output of a 5V DAC.

An Automatic Level Control circuit keeps the vertical height constant as the vertical scan rate is changed (in a multi-sync monitor for example), or if the supply voltage or temperature changes. The filter for the ALC is connected to Pin 18 and uses a capacitor that is 20X the value of the oscillator capacitor.

To allow rapid or dynamic adjustment of the vertical height, the ramp is passed through a second gain controlled stage before reaching the output. A voltage between 3V and 4V on Pin 6 will control the height over a 1 to 1.3 range. After the height has been set, the ramp voltage is converted to a differential current by a resistor at Pins 21 & 22. A 10Kohm resistor will produce at least 1mA differential current superimposed on a .3mA dc current. Because the voltage to current conversion follows the ALC circuit, to maintain the amplitude constant with temperature changes, a carbon film resistor is recommended and should be placed as close as possible to the I/C.

LM1295 DC Controlled GEOMETRY CORRECTION



35

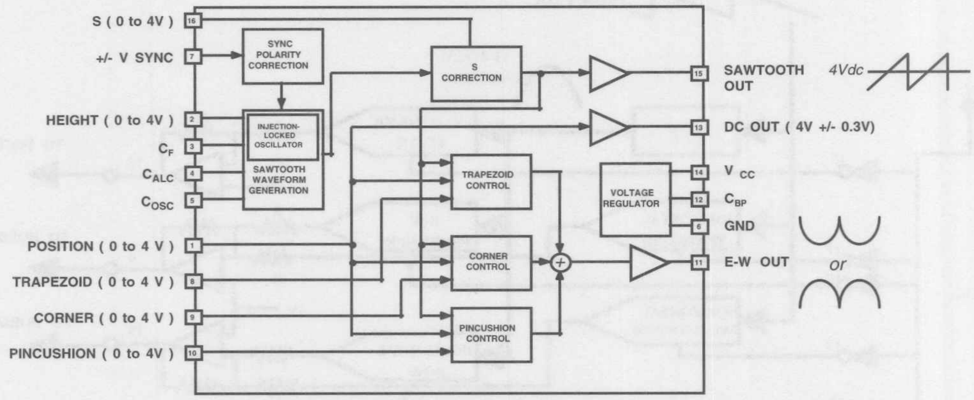
The voltage ramp from the oscillator section is applied to a parabolic function generator and two multiplier circuits which are being used as gain controlled amplifiers. The other inputs to the multipliers are dc voltages. AT 2V, the multiplier output ramp is zero. From 2V to 0V an increasing positive ramp is obtained. From 2V to 4V an increasing negative ramp is obtained. Similar dc control determines the polarity and amplitude of the three multipliers processing the output of the vertical parabola generator.

The vertical parabola is buffered to Pin 16 to provide a dynamic focus control voltage.

Summing the parabola with the Horizontal Trapezoid in a ratio of 3:1 gives the pincushion and keystone correction waveform at Pin 9. This waveform is used to dynamically change the horizontal scan width.

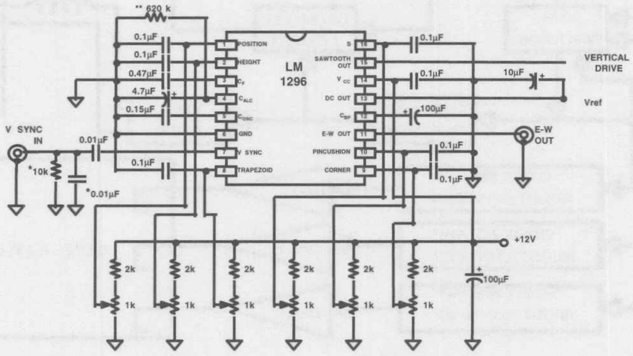
Finally summing the parabola with the output from the Horizontal Parrallogram multiplier will generate a voltage at Pin14 to correct the horizontal centre position (phase).

LM1296 BLOCK DIAGRAM



The LM1296 is also designed to provide geometry correction, but in a 16 pin package. It has the same major features as the LM1295; Injection locked oscillator with dc controlled height adjustment, pincushion and trapezoid correction. A single ended vertical drive waveform is provided.

LM1296 APPLICATION CIRCUIT



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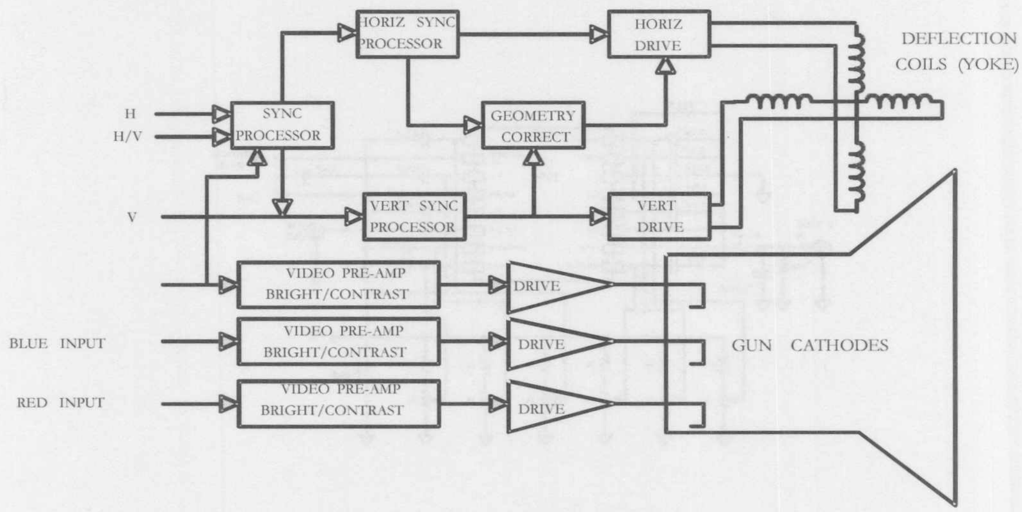
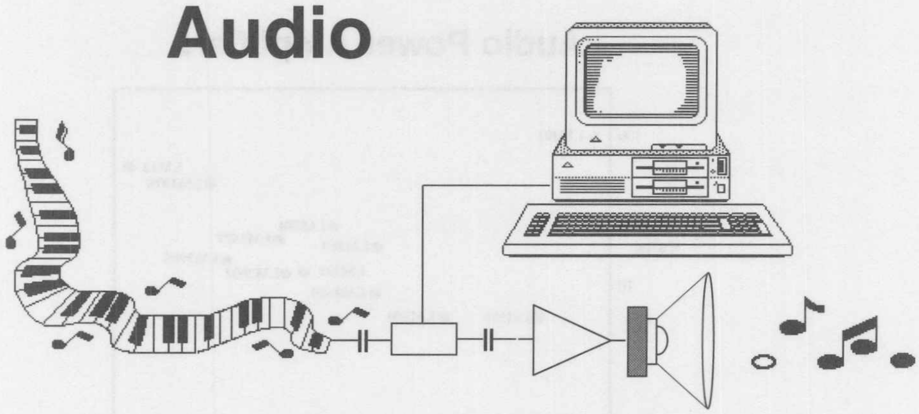
With the fewer pin connections, the LM1296 requires fewer external components than the LM1295, and, as shown here, potentiometers can be used to provide the correction control voltages. The LM1296 can accept either polarity of vertical sync, and the S correction is done internally.

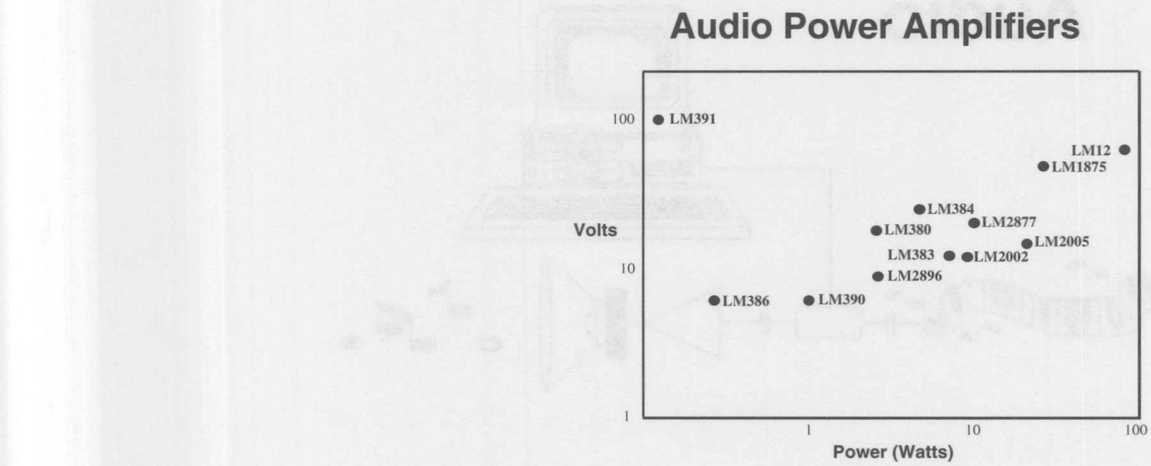


38

National can provide most of the components that are needed to build high quality VGA and SVGA monitors. The high level of integration causes a dramatic decrease in the number of components and the size of the circuit board. In particular, this makes neck mounted boards feasible, thus reducing lead lengths that carry easily degraded high frequency signals and contribute to emi problems. Dc controls in the 0V to 4V range make implementation of automatic alignment easy and effective.

Audio





40

National has been in the monolithic Audio Power Amplifier business for over two decades, supplying devices for a variety of applications, from Pachinko machines to Automotive, portable audio to home stereos, and many more. These amplifiers have been produced as singles and duals, low power and medium power, with distortion specs that range from 10% THD down to 0.01% THD.

Less than 10 years ago, high power meant an amplifier capable of delivering 20 Watts or more, and this part of the market, relatively small in I/C volume terms, was dominated by discrete transistor or Hybrid designs. Usually they were expensive to manufacture and offered little in terms of the self protection features that are virtually standard with monolithic devices. Today, new monolithic power amps are displacing these designs, with power levels approaching or exceeding 100 Watts.

At the other end of the power spectrum, interest in multi-media PC applications has generated a demand for quality Audio amplifiers that can be integrated into the PC environment. Here the challenge has been to develop power into 8 ohm speakers with only 5 Volt power sources available.

In this next section we will see how National is providing solutions for both High Power Audio and delivering audio power from 5 Volt supplies.

AMPLIFIER POWER RATINGS

1) AVERAGE POWER = 50 Watts

2) CONTINUOUS RMS POWER = 50 Watts

3) MUSIC POWER = 65 Watts

4) PEAK POWER = 100 Watts

5) POWER = 50-100Watts

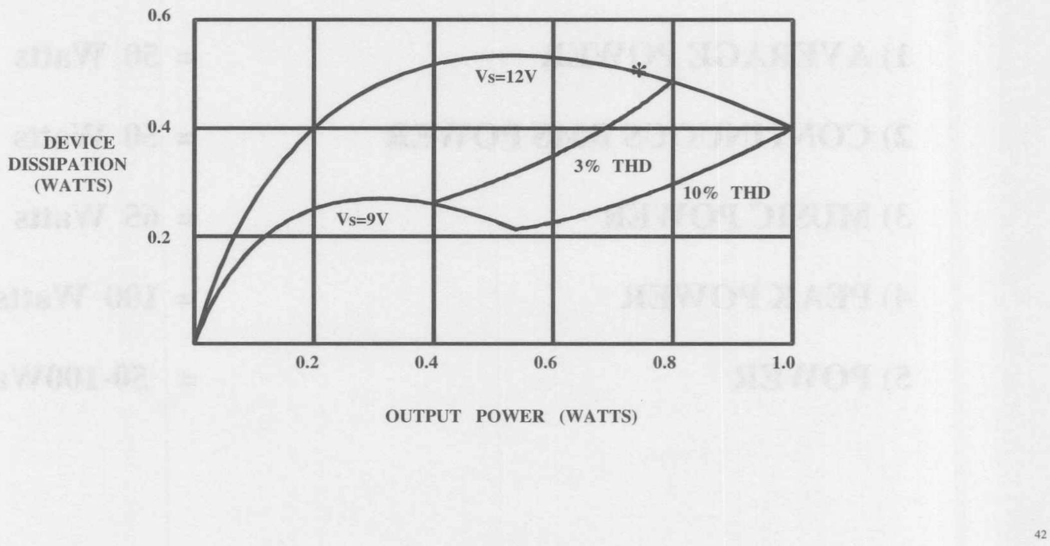
41

Power ratings can be confusing, and this is not always accidental. To begin with, the unit of power that is used, the WATT, is not the simple product of a DC current and DC voltage applied to resistive load. Nor is the actual load (a speaker) truly resistive.

Manufacturers have used various ways to express the power capability of the audio amplifier, usually with the goal of putting their product in the best light. Typical power ratings are 1) Average Power, 2) Continuous RMS Power, and 3) Music Power. Actually the first two are the same and are the measure of power using rms values for the sine wave current in a resistive load and the sine wave voltage applied across it for a sustained period of time. Music Power also uses rms values in calculating the power but in the actual end product the power supply voltage will depend on the amount of power drawn. The power rating is calculated before the supply has an opportunity to sag. Therefore more voltage is available than if the amplifier had to deliver sustained power. This means our 50 Watt continuous amp can be rated at 65 Watts or more! The fourth rating, Peak Power, is obtained by multiplying the peak voltage across the load, by the current drawn when that peak voltage is applied. This 'instantaneous' power calculation raises the power of our amplifier to a whopping 100 Watts. A simple 'Power' rating could be any of the above.

The moral: when comparing amplifier specifications, compare the way they are measured.

POWER vs DISTORTION



Another part of the comparison process is to note the distortion level at which the power rating is determined. On many amplifier data sheets the Typical Performance Curves section will show what this means. For example, most of the smaller power amplifiers are rated at 10% THD. The Typical Curves show why. At 10% THD the power output is significantly higher than at 3% THD or less.

An LM386 will deliver about 0.75 Watts into a load without distortion. The output waveform will just begin to show clipping at 3% THD, and by the time 5% THD is reached the power is up to 0.9 Watts, a 20% increase. Allowing the THD to reach 10% increases the rated power to 1 Watt, 30% more than the undistorted power level.

In the section that follows, power ratings will always be given as the Continous RMS Power rating measured at <1% THD unless otherwise noted.

A Revolutionary PC Audio Power Amplifier...

BOOMER™

Goal: To obtain high fidelity output power for PC Audio Applications

Design criteria are: 1 watt into 8 ohms with less than 1% Total Harmonic Distortion from a 5V power supply in a SO package

43

One of the recent developments in power amplifiers is the need to improve the speaker driver (and the speaker!) in PCs. If the internal 5 Volt power supply is the only high current power source available, putting power into a quality 8 ohm speaker represents somewhat of a challenge.

National has recently introduced a new range of audio power amplifiers designed to deliver the most power on limited power supply sources. This family of parts has been named BOOMER™, in part because of the surprising amount of power they can deliver into standard 8 ohm speakers. The first two devices are the LM4860 and LM4861.

Evolution of Boomer

Design Specifics...

- 1 watt into 8 ohms requires a voltage swing of 8 V(p-p) and a current of 353 mA rms from the amplifier
- The challenge - 8 V(p-p) from a 5 V supply

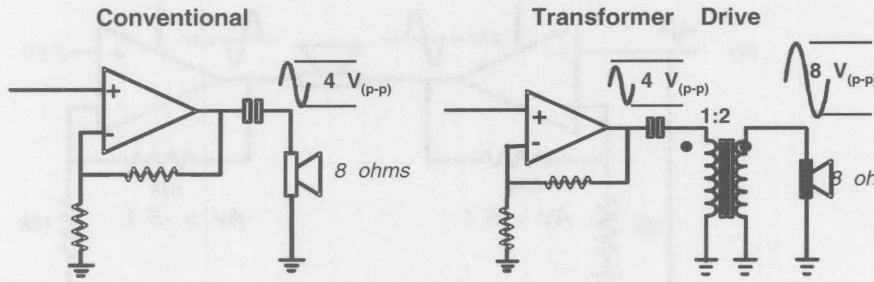
44

To meet the goal of producing 1 Watt into an 8 ohm load from a 5 Volt supply means discarding the conventional audio amplifier topology. A simple calculation shows why this is the case. A 1 Watt power level is represented by an rms current of 355 mA in an 8 ohm load. To produce this current the voltage applied across the load must be 2.8 Vrms. This means that the amplifier output must swing at least 8 V(p-p), which clearly exceeds the power supply voltage of 5 Volts.

Alternatives, such as using transformers on the amplifier output, or providing a stepped up power supply voltage, were discarded since another goal was to use the least number of external components. This leads to the use of a bridge topology, which not only can provide the required power level, but also yields a high level of integration with minimal external components. As shown above, the output power equation is no longer $V^2/8R$, where V is the peak-to-peak swing across the load R, but is now $V^2/2R$. To achieve 1 Watt in 8 ohms now requires 4 Vp-p.

Bridge amplifiers are not new, but not everyone is familiar with the way they work. In this instance we are using the bridge because the supply voltage is limited. Early bridge audio amplifiers were developed to eliminate audio transformers.

Increasing the load POWER



Power = $(4)^2/64$ = 0.25 Watts

Power = $(8)^2/64$ = 1.0 Watts

45

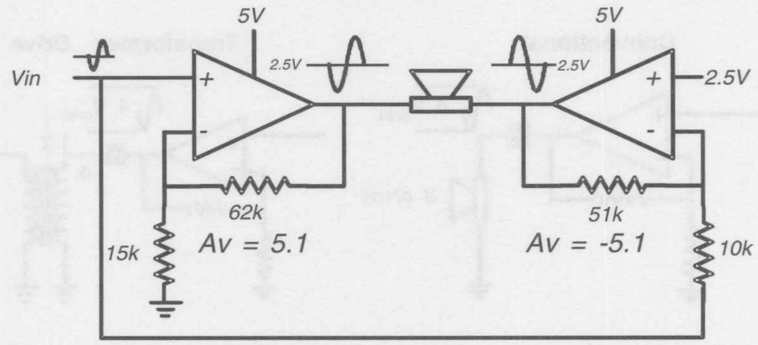
The starting point to understanding the power capabilities of bridge amplifiers, whether from an intuitive or mathematical viewpoint, is to look at how a bridge amplifier works. These amplifiers are also known as BTL amplifiers, or **B***ridged* **T***ransformer***L***ess* amplifiers, since they were originally developed to replace transformer coupled loads.

The transformer turns ratio allowed the *AC* signal from the amplifier to be stepped up to a larger signal across the load, which was particularly important in the early days when transistor breakdown voltages restricted the size of the supply voltage, which in turn limited the voltage swing from the amplifier and the amount of power that could be delivered to the load. For example, with a 6 Volt rating on a transistor, the output stage could not be designed to deliver much more than about 4 V(p-p). In 8 ohm loads this amounts to an rms power level of only 0.25 Watts. Placing a transformer between the output stage and the load enables the output voltage swing across the load to be larger than the transistor rating.

As shown here, a 2:1 turns ratio (the secondary has twice the turns of the primary), will increase the load swing from 4 V(p-p) to 8 V(p-p). If the transistor can deliver the extra current then the power is increased to 1 Watt.

The BTL design allows the output voltage swing to be doubled without using a transformer, with the same fourfold increase in power delivered to the load. Today, the incentive for using a BTL design is not the breakdown capability of the amplifier, but to increase the output power capability on the low system supply voltages that are now being used.

The Basic Bridge Amplifier



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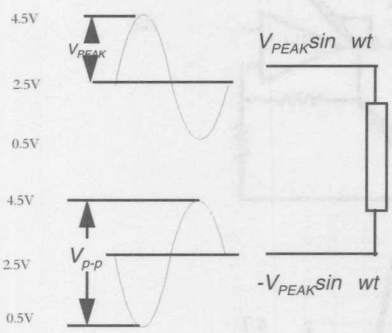
There are a number of different ways of connecting bridge amplifiers, but all the topologies have one thing in common. Two amplifiers are used, driving opposite ends of the load, with the output from one amplier being in anti-phase to the output of the other amplifier. Usually both amplifiers have the same output amplitude (to maximise the signal swing and produce symmetrical clipping) but this is not a requirement. What *is* important, is that for the signal frequency range of interest, as one amplifier swings in the positive direction, the other amplifier swings in the negative direction (or vice-versa).

In the absence of the ac signal input, both amplifiers have their outputs biassed halfway between the supply rails. For a single 5 Vdc supply this means both outputs are at 2.5 Vdc. If the outputs are well matched there will be no DC current flow in the load. Notice this means that the large speaker coupling capacitors have been eliminated.

Each amplifier has the same gain (X5.1), set by the ratio of the feedback resistors from the output back to the inverting input. For the left-hand amplifier the signal gain (non-inverting) is (1 +62k/15k) = 5.1. For the right-hand amplifier it is -5.1k/10k = -5.1. For the moment, we have neglected biassing or stabilisation components.

Putting FOUR TIMES the POWER into the Load



$$V_{LOAD} = (V_{PEAK}\sin wt) - (-V_{PEAK}\sin wt)$$

$$= 2\ V_{PEAK}\sin wt$$

$$V_{RMS} = \frac{2\ V_{PEAK}}{\sqrt{2}} = \frac{V_{p\text{-}p}}{\sqrt{2}}$$

$$\text{Therefore Load Power (Bridge)} = \frac{(V_{p\text{-}p})^2}{2R}$$

If $V_{p\text{-}p} = 4V$, then Po = 1 Watt

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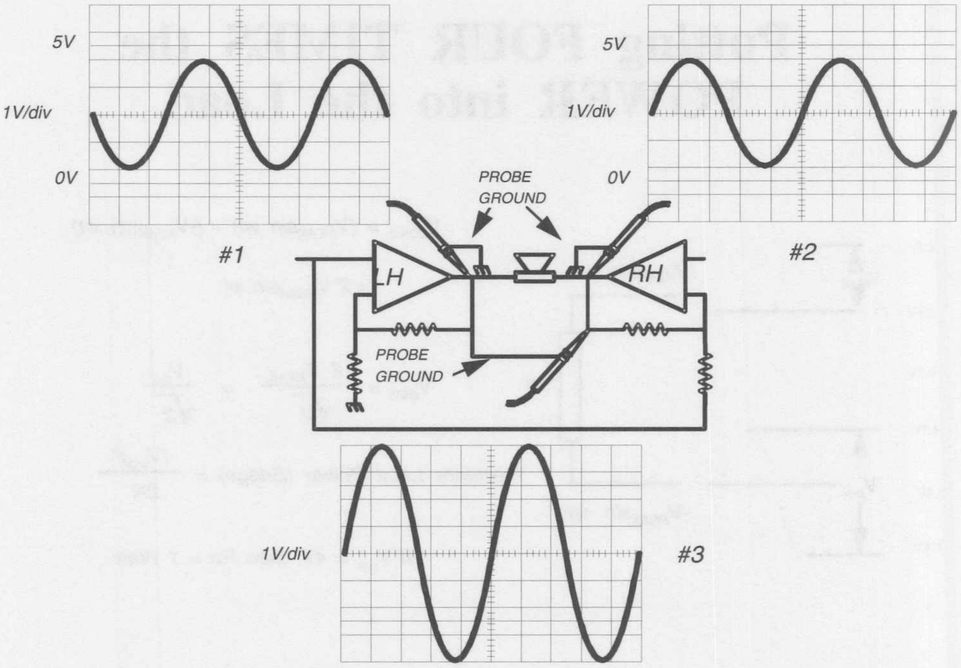
Since signals are applied to each end of the load, the actual load voltage is the difference between the signals, or the *differential voltage*. In the case where the signals have the same amplitude (V_{PEAK}), but are opposite in phase, the peak load voltage is $2V_{PEAK}$.

Let's follow the voltages at each end of the load through a complete cycle of the drive waveform. As the upper end of the load moves from 2.5V to 4.5V, the lower end of the load moves from 2.5V to 0.5V for a peak change of 4V, ie $2V_{PEAK}$. During the next part of the cycle both ends of the load are returned to 2.5V. Next the upper end of the load moves down to 0.5V while lower end moves up to 4.5V for another peak change of 4V. In the last part of the cycle both ends of the load return to 2.5V. The total change through a complete cycle is $8V_{p\text{-}p}$. Since the load power is proportional to the square of the voltage, the power developed by the bridge amplifier is *four times* the power developed by a single ended amplifier with the same voltage swing (V_{PEAK}).

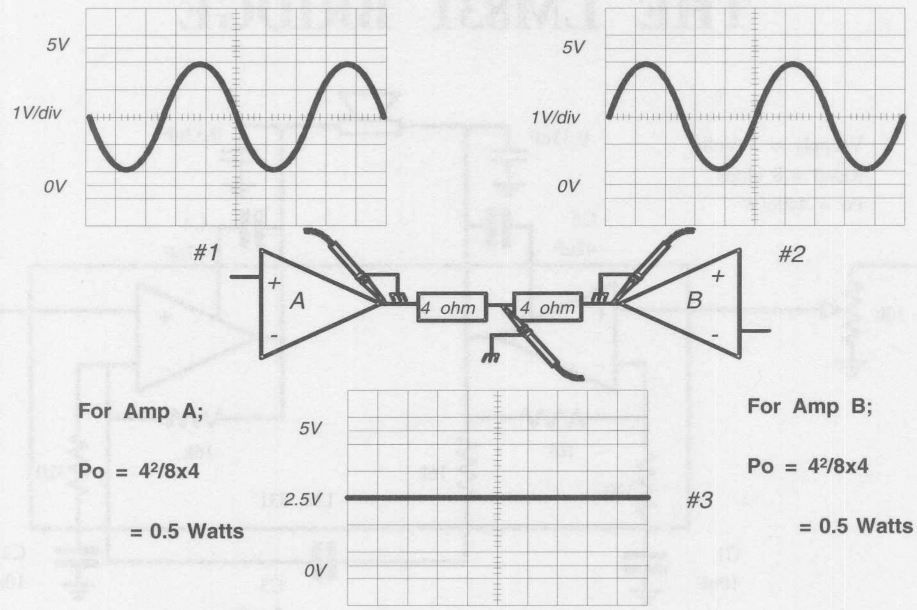
If *each* amplifier is swinging the same 4V(p-p) output as the single-ended amplifier, and the load is still 8 ohms, then the load voltage swing is $8V_{p\text{-}p}$ and the load power is $8^2/8 \times 8 = 1$ Watt.

Most of the time bridge amplifiers are operated with each amplifier having an identical swing capability. Since it is easier to measure the swing from an amplifier output to ground rather than differentially across the load, for bridge amplifiers the output power equation is modified as shown above, where $V_{p\text{-}p}$ is the peak to peak voltage swing of *one* of the amplifiers.



O.K, seeing is believing? If we look at the waveforms at the amplifier outputs what will we see? Here we are showing the oscilloscope traces that will be observed when the 'scope probe is attached to the amplifiers at the points shown above. For the LH amplifier the output is 4V(p-p), photo #1. Notice that the 'scope probe ground (attached to the body of the probe) is connected to the same ground used by the amplifiers. Moving the 'probe to the RH amplifier, we again see 4V(p-p), photo #2......and if the 'scope is properly synchronised to the *signal source* we will note that this waveform is 180° out of phase with respect to the waveform at the LH amplifier. What we are looking in each case is the AC voltage swing at one end of the load with respect to ground. To see the total swing across the load, connect the probe to one end of the load and lift the probe ground from the circuit ground and connect it to the other end of the load*. This is photo #3 which shows 8V(p-p) across 8 ohms. Notice that for photo #3 there is no longer a DC reference for the signal since the pobe is "floating" with respect to the circuit ground.

8V(p-p) across an 8 ohm load gives a load power of 1 Watt (8^2 /8 x 8).



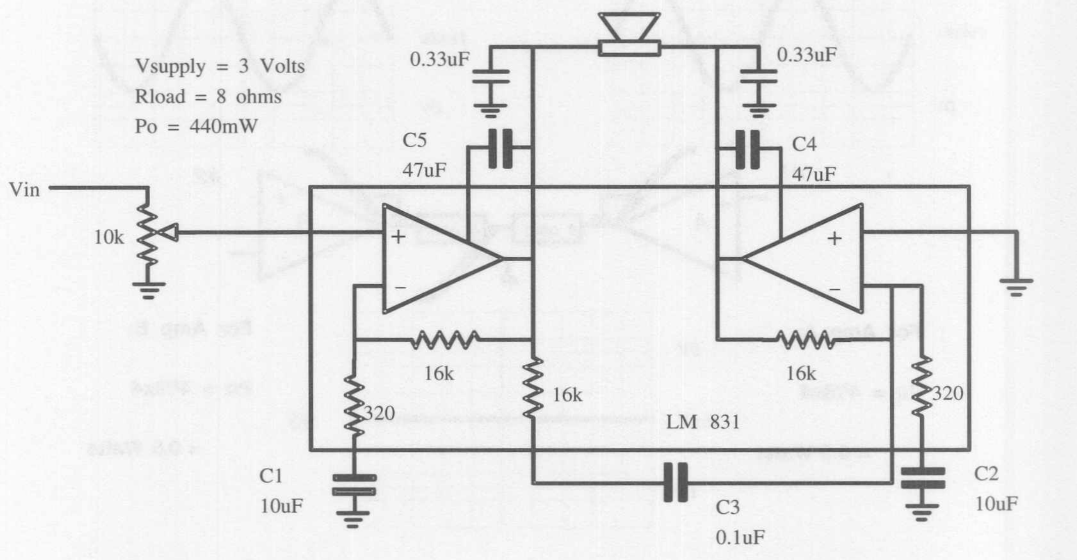
The Intuitive approach:

While you can't disbelieve the 'scope photos, differential readings are not very intuitive for most people, particularly in a single supply voltage situation where '*ground is ground*'. Having the 'scope probe 'ground' move during the measurement can be uncomfortable. Here's another way of looking at it.

Replace the 8 ohm load with two 4 ohm loads connected in series. If the output of Amplifier A is 2.5 Vdc and the output of Amplifier B is also 2.5 Vdc, then the centre tap between the two loads is also 2.5 Vdc. Connect the 'scope probe to this point. When Amplifier A output goes up to 4.5 V, Amplifier B output will go down to 0.5 V. The centre tap will stay at 2.5Vdc. When Amplifier A swings down to 0.5V, Amplifier B will swing up to 4.5 V. The centre tap will still be 2.5Vdc. As shown in photo #3, the centre tap will stay at 2.5Vdc regardless of the signal swing (if the amplifier outputs are in anti-phase over the frequency range of the input signal).

If amplifier A swings from 0.5V to 4.5V or 4V(p-p) with respect to the centre tap, across 4 ohms, the power delivered is $(4)^2/8 \times 4 = 0.5$ Watts. Amplifier B is simultaneously swinging 4.5V to 0.5V or 4V(p-p) across the other 4 ohm load for another 0.5 Watts. The total power delivered (into 8 ohms) is the sum of the power delivered into each 4 ohm load, ie 0.5W + 0.5W = 1Watt

THE LM831 BRIDGE



The LM831 is a dual monolithic audio amplifier that has been optimised for low power supply voltage operation. Each amplifier is able to drive a 4 ohm load and having both amplifiers in one package makes it an ideal candidate for bridge applications.

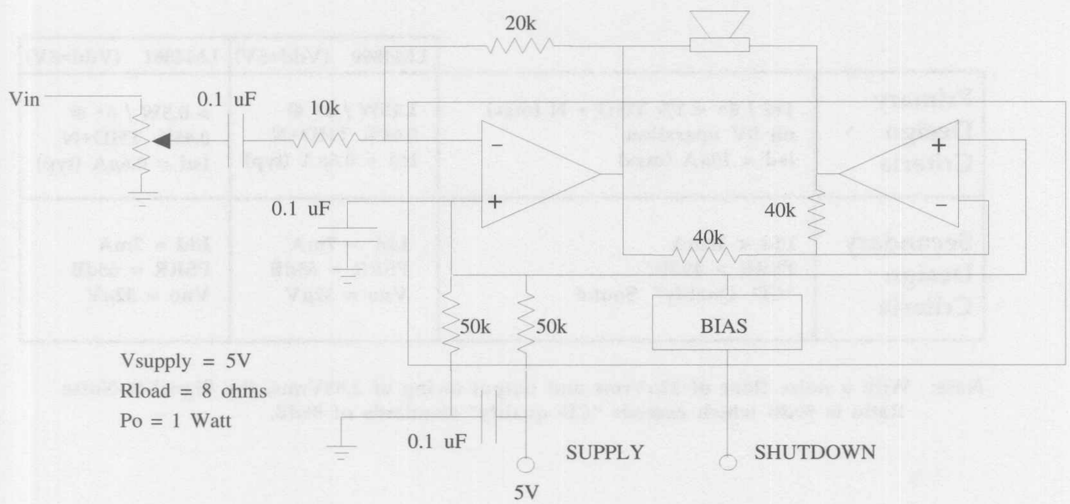
Although the signal is directly connected to the non-inverting amplifier, the signal input to the inverting amplifier is obtained from the output of this non-inverting amplifier. Since the signal gain has already been raised by the non-inverting stage, the inverting amplifier operates with unity gain (0dB) so that the same signal swing is obtained on each side of the load.

In more detail, the *signal* gain of the non-inverting stage is set by (16K + 320)/320 = 51 or 34dB. The non-inverting amplifier output is the signal source , through 16 Kohm, to the inverting amplifier, which has another 16 Kohm feedback resistor from its output back to the inverting input. Therefore the gain of the inverting amplifier is set by the ratio of these resistors, -16K/16K = -1 or 0db. Since the input signal level to the inverting amplifier is the output signal level from the non-inverting stage, a gain of -1 means that the same signal level is applied to the other side of the load, simply reversed in phase.

The LM831 was designed with low, single supply voltages in mind and, as such, has several design features to make this possible. In common with many bipolar single supply amps, PNP transistors are used at the input stage. This allows the source DC voltage to be at ground potential. The gain setting resistors are included within the package to minimise the external components required and an internal bias circuit sets the output DC voltage at half supply to maximise the output voltage swing.

Even with internal gain set and bias, a number of external components are still required.

LM4860-BOOMER^{R}



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Where maximum power with minimum external components is the goal, a slightly different approach can be taken.

This is BOOMER^{R}, designed to deliver over 1 Watt into 8 ohms from a 5 Volt supply. Notice that this bridge circuit has a different topology than that previously described, and that there are far fewer external components. The reduction in external components is due primarily to the technology that is used.

Putting both amplifiers in a single package, and configuring the gain for a bridge application eliminates the load coupling capacitor. The gain stage is an inverting amplifier which becomes the source for a unity gain inverting follower amplifier. Because the LM4860 is built on a CMOS process, the follower amplifier can be internally configured for unity gain without risking stability, yet still have sufficient bandwidth to ensure low distortion at high signal frequencies. Using the inverting amplifier as the gain stage enables external resistors to set the gain without requiring large decoupling capacitors in the feedback network to preserve the low frequency response. Because both the inputs and the outputs of each amplifier are DC biassed at half supply voltage, the signal connection between the amplifiers is DC, eliminating a large coupling capacitor. Even if the source impedance is high enough to affect the closed loop gain of the inverting amplifier, since the follower amplifier is driven from the inverting output, the clipping performance for both sides will remain symmetrical. With a CMOS process, the complementary nature of the output stage devices mean that snubber networks are not required on the outputs to prevent the "bottom side fuzzies". This CMOS output stage can also swing closer to the supply rails than a conventional bipolar stage, maximising the output voltage swing into the load (rail to rail performance). This removes the need for *bootstrap* capacitors.

The end result is an amplifier with remarkable performance on a low 5 Volt supply. The BOOMER^{R} LM 4860 will deliver over a Watt into 8 ohms at less than 1% THD.

LM4860/4861 PERFORMANCE

		LM4860 (Vdd=5V)	LM4861 (Vdd=5V)
Primary Design Criteria	1W / 8• < 1% THD + N (max) on 5V operation Isd < 10μA (max)	1.15W / 8• @ 0.68% THD+N Isd = 0.6μA (typ)	> 0.5W / 8• @ 0.45% THD+N Isd = 0.6μA (typ)
Secondary Design Criteria	Idd < 10mA PSRR > 40dB "CD Quality" Sound	Idd = 7mA PSRR = 65dB Vno = 32μV	Idd = 7mA PSRR = 65dB Vno = 32μV

Note: With a noise floor of 32μVrms and output swing of 2.83Vrms, the Signal-to-Noise Ratio is 99dB which exceeds "CD quality" standards of 96dB.

As can be seen, using the bridge topology results in good performance on low supply voltages. Bridge amplifiers can also be used for much higher power levels. In fact, as we will see, monolithic power amplifiers are no longer limited by their current/voltage capabilities, but more by the power dissipation capability of the package. Next we will take a look at National's new high power audio amplifiers to see why this is the case, and how a bridge topology can overcome the heat dissipation limit to output power.

Power Audio Amplifiers

LM2876	LM3875/LM3876	LM3886
25W/8ohm Continuous	40W/8ohm Continuous	60W/4ohm Continuous

For some time monolithic Audio Power Amplifiers have been limited in power output capability to about 20 Watts rms. At this level they have been able to provide Hi-Fi specifications and include many protection features not commonly found in discrete or Hybrid designs. It is the ability to include such features with a relatively low penalty in circuit complexity or cost that make the integrated circuit amps very attractive to manufacturers of audio equipment.

Now, with the introduction of the Overture™ series of Audio Power Amplifiers, National is raising the power levels at which these features are included, yet still providing Hi-Fi performance specifications.

The Overture series uses an advanced protection scheme, developed from the LM12, which allows superior protection at very high power output levels. The LM2876 can deliver 25 Watts into 8 ohm loads, the LM3875/3876 can deliver 40 Watts into 8 ohms, and the LM3886 is able to deliver 60 Watts into 4 ohms.

The most highly integrated & dynamically protected high performance Audio Amplifiers in the 25W to 60W range

SPiKe™ Protection

- *Overvoltage Protection.*
- *Undervoltage Protection.*
- *Short Circuit Protection.*
- *Thermal Runaway Protection.*
- *Instantaneous Peak Temperature Protection.*

Highest Performance:

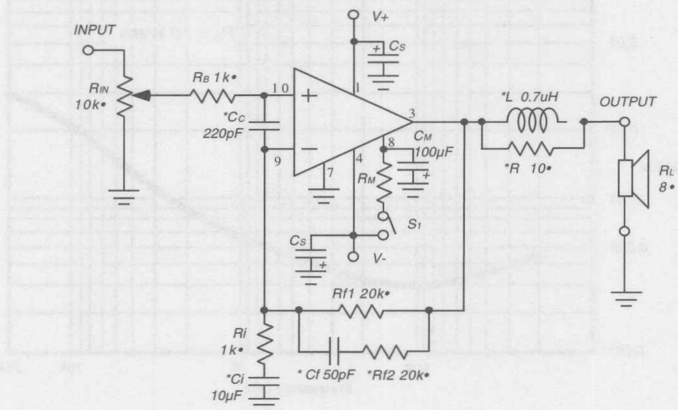
- *60W into 4ohm at less than 0.05% THD+N from 20Hz-20kHz.*
- *Low Noise Floor of 2μV (input referred).*
- *High Slew Rate of 11Vμs*
- *High PSSR of 120 dB.*

The amplifiers in this series are designed to operate from high supply voltages, as much as 90 Volts, and deliver peak currents to the load that can exceed 10 Amps. The high voltage ratings mean that the amplifiers can withstand high-line conditions yet still be able to deliver the rated power with a nominal supply. They also have an undervoltage lockout feature. When the power supply is first turned on, the output stage is disabled until there is sufficient supply voltage to ensure proper biassing of the internal circuits. When the output stage turns on, it does so without causing a large transition to either supply rail at the output. On a split supply the output voltage is at 0 V DC until a signal is applied. This provides pop-free operation.

If the amplifiers are used without completely adequate heatsinking, thermal shutdown circuits are activated to prevent the die temperature exceeding a preset internal limit. Also instantaneous hot spots are monitored with the SPiKe™ protection circuit, preventing destruction of the output devices when the SOA is exceeded.

Many of these features are not available in discrete designs at comparable power levels, yet the performance of the monolithic devices matches the best discrete or Hybrid designs.

LM3886 Split Supply Application Circuit



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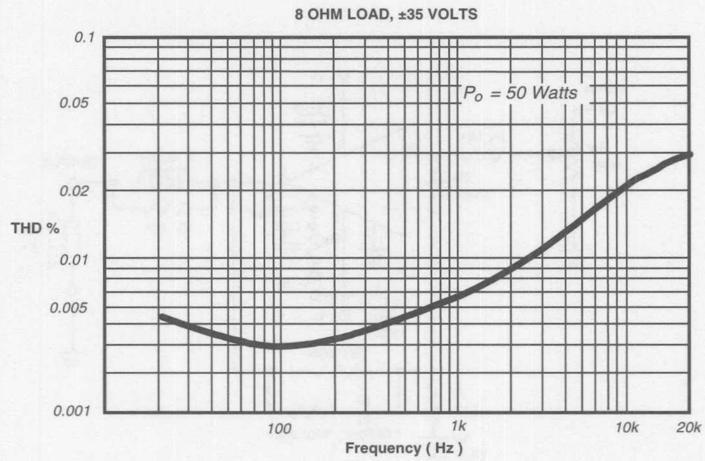
This is the complete schematic of a power audio amplifier using the LM3886. The gain is set by the ratio of the resistors Rf1/Ri and the input impedance is set by the 10kohm volume potentiometer. As with any high power amplifier, the power grounds should be separated from the signal grounds, and good supply decoupling used close to the device supply pins.

LM3886 THD vs Frequency



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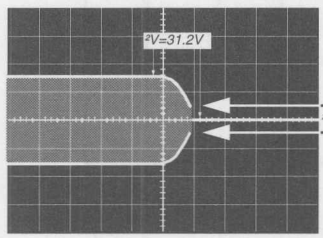
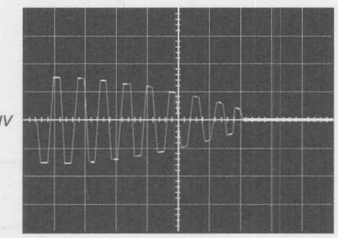
The performance of this amplifier at the 50 Watt power output level is very good. The distortion at 1kHz is only 0.005 %. The distortion will increase at very low frequencies from thermal effects in the output devices causing offsets at the input stages. At very high frequencies the distortion increases because there is less open loop gain at high frequencies.

Undervoltage Protection





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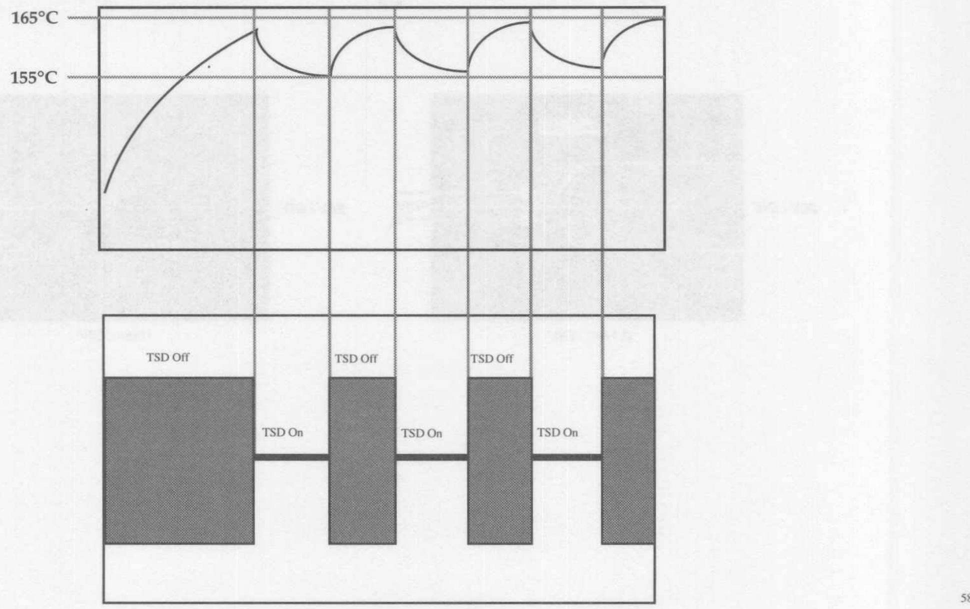
When the voltages applied to the amplifier are excessive, the predominant issue is survival. When the voltages are much less than required for rated performance, the issue is a graceful withdrawal from operation. Once the supply voltage gets too low, many amplifiers will cause the output to slam against one of the supply rails rather than remain centred around O VDC. This produces a disconcerting pop from the speakers. While not necessarily damaging, turn-on and turn-off pops are not popular with the majority of listeners, and to avoid this problem many amplifiers include relays that disconnect the load (ie the speakers) when the supply is turned on or off.

The Overture series incorporate an undervoltage lock-out system that prevents the output stage from actively supplying the load unless the supply voltage is above a specified minimum level. As shown here, as the power supply voltage falls the output signal decreases. The signal is clipped because the input signal is kept constant at a level to produce the full voltage rated output. However, once the supply drops below 18V, the output stage is disconnected from the load so that the load voltage drops immediately to 0 VDC. There are no transient swings to either supply rail.

Thermal Runaway Protection



The thermal protection circuit monitors the whole die temperature and is (relatively) slow in operation. It takes many cycles of the signal to raise the temperature, and many cycles before normal operation is restored.

The top waveform depicts the temperature rise and fall of the junction as the Thermal Shutdown Circuitry is activated.

If proper heatsinking is not utilized, the die will heat up due to the poor dissipation of power. Once the die reaches its maximum temperature rating, Thermal Shutdown Protection is enabled, thus regulating the temperature of the die.

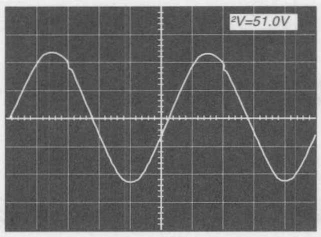
Once the die temperature reaches approximately 165°C the output is driven to ground, as shown in the lower waveform. This is likely to cause a pseudo "pop" at the output since the waveform being amplified at the time is cut off and driven to ground.

The intention of the protection circuitry is to prevent the device from being subjected to unbearable temperatures. It is this schmitt trigger effect that prolongs the life of the device and provides higher reliability to the customer.

SPiKe™ Protection Response With Increasing Case Temperature

Tc = 75.6°C

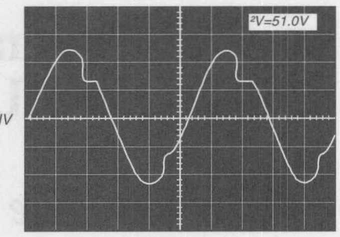
10V / DIV



2msec / DIV

Tc = 85.0°C

10V / DIV



2msec / DIV

SPiKe™ is a National innovation, pioneered with the LM12, that provides a way to monitor any spot on the output transistor that is experiencing higher temperatures than the rest of the transistor. Normally a hot spot will cause a decrease in the V_{BE} voltage at that region (for silicon the V_{BE} falls -2mV/°C). A lower V_{BE} means that more current will flow in that region raising the power dissipation....and causing the V_{BE} to fall even more. Inevitably the heat created will cause a failure that destroys the transistor. To prevent this, SPiKe monitors the temperature and shuts down the drive if it gets too high. Unlike the thermal shutdown circuit, which is monitoring the entire die, SPiKe is looking only in the region that is getting hotter than the surrounding die. Response is virtually instantaneous.

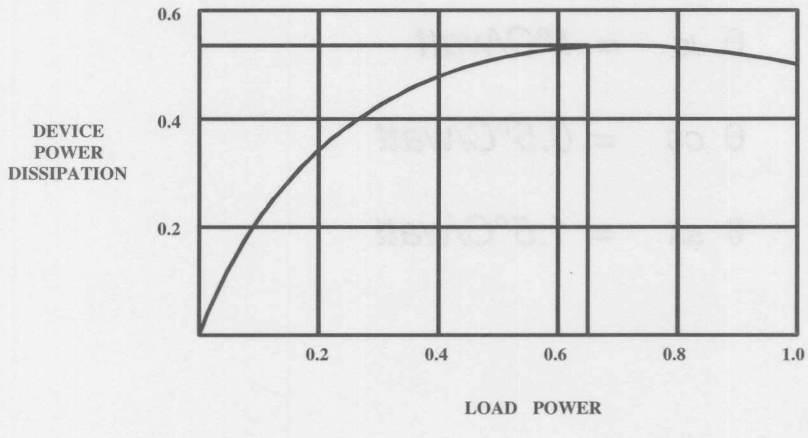
How much Power can the LM3886 deliver?

What size Heatsink is needed?

In discussing protection circuits, particularly thermal overload protection circuits, it might be thought that these are included to accomodate unusual circumstances. Unfortunately, for audio power amplifiers in particular, this is not true. Many power circuits are designed on the basis of the voltage and current required to generate a specific power level in the load. What is forgotten is that a substantial amount of the power available from the power supply is not delivered to the load. Instead it is dissipated in the amplifier....as heat! A proper design will include a heatsink to remove this power, but when the heatsink is inadequate, the temperature goes up. Routine activation of the thermal shutdown will save the amplifier, but not the patience of the listener.

The question arises, how much heatsink is required--or even possible? Also, at some point the practical limit on heatsink size will be reached. Then the question becomes, how much power can a monolithic audio amplifier deliver?

POWER vs DISTORTION



This is the same type of curve that we showed previously, for the LM386. Both amplifiers are operated in Class B, and the device dissipation versus power output follows this classic shape. However we have normalised the power output to unity at the maximum ***undistorted*** load power. Note that the maximum device power dissipation occurs at about 65% of the maximum load power. Therefore if we calculate the maximum power dissipation of the amplifier package and heatsink combination for a given ambient temperature, we can determine the maximum ***undistorted*** load power that the amplifier can deliver, irrespective of the current and voltage ratings of the device. This will be the maximum allowable power dissipation Pd_{MAX} , divided by 0.65.

Power dissipation is calculated from the package thermal resistance, the heatsink and interface thermal resistance, the maximum allowable device junction temperature and the maximum operating ambient temperature.

$$\theta_{JA} = 37°C/watt$$

$$\theta_{JC} = 1°C/watt$$

$$\theta_{CS} = 0.5°C/watt$$

$$\theta_{SA} = 1.5°C/watt$$

62

In this case the amplifier comes in a TO-220 style package. Without a heatsink, the thermal resistance, junction to case, is around 37 °C/W. What that number means is that for every watt dissipated by the amplifier, the junction temperature will increase by 37°C. We are more interested in the thermal resistance, junction to sink. These amplifiers have very large die sizes and the thermal resistance, junction to case, approaches that of large discrete transistors in T0-3 packages, ie about 1^0C/W. The case to heatsink interface will add another 0.5^0C/W, which leaves us with the size of the heatsink. There are no hard and fast rules here. The heatsink size can be designed to keep the device out of thermal shutdown at very high ambient temperatures.....75^0C for example. Alternatively smaller sinks can be used for lower rated temperatures on the assumption that the heatsink thermal mass will allow the device to ride out periods of high power dissipation. Apart from the economic considerations of a large heatsink, at some point the disparity in size between the heatsink and the package will prohibit the uniform transfer of heat. For these reasons we will consider that a heatsink thermal resistance of 1.5^0C/W is about as large as we can use. Now let's consider a maximum operating ambient temperature of 65^0C.

Output Power Depends On The HEAT SINK

$$\text{If } \theta_{ja} = \theta_{jc} + \theta_{cs} + \theta_{sa} = 3°\text{C/Watt}$$

At Ta = 65°C

$$Pd_{(max)} = (165 - 65)/3 = 33.3 \text{ Watts}$$

$$Po = 1.5 \times Pd_{(max)} = 50 \text{ Watts}$$

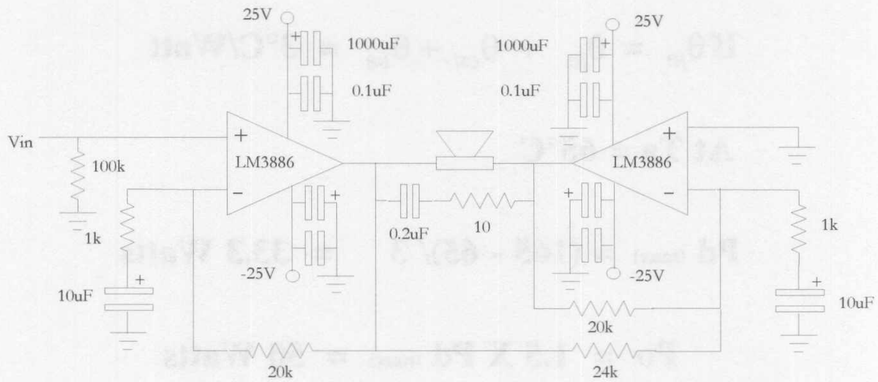
63

With the total circuit thermal resistance, junction to air, of 3°C/W, then at 65°C ambient, the maximum device dissipation will be 33 watts. If the supply voltage rails are chosen such that this limit is not exceeded (power dissipation is dependent on the supply voltage) then the maximum power output *before distortion* is just over 50Watts..

Lower ambient temperature ratings, and/or even larger heatsinks, will allow bigger continuous power ratings, but probably not much above 60 Watts for a single package amplifier. Notice that we have not considered the electrical specifications in arriving at this result. The maximum power is limited by package thermal characteristics at these output power levels.

The most practical way of increasing the power is to use more packages (much as a discrete transistor design does). A bridge can increase the available power to over 100 Watts since the power dissipation is shared between two packages. Obviously the heatsink is twice as large.

High Power Bridges



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Clearly, at the 100 Watt power level, very close attention must be paid to minimising the total thermal resistance. Higher power levels may be attainable with lower operating ambient temperatures, better (ie bigger) heatsinks and minimising the effect of the insulating washers. The insulating washer can be eliminated by using electrically isolating thermal grease (micro-balls of glass mixed into the grease) or overmoulding the tab of the power package. Leaving the heatsink at a high dc potential is *not* recommended, nor is using a single ended power supply unless some provision is made for eliminating turn-on pops.

AUDIO AMPLIFIERS FOR BRIDGE APPLICATIONS

Duals (1 I/C Package)	Singles (2 required)
LM831[1]	LM380
LM1877/2877[2]	LM383
LM1896/2896[1]	LM384
LM2878/2879[2]	LM386
LM4860/4861[1]	LM388
LM1876	LM390
	LM1875
	LM2876
	LM3875/76/86

[1] Characterised in Bridge mode

[2] 16 ohm loads only

65

Most audio power amplifiers are specified for single-ended operation. In the rare instance a bridge circuit is shown, often this demonstrates only the possibility. Both previous examples have used I/Cs where the data sheet not only shows bridge applications, but additional specifications are given for the bridged application. All of National's audio amplifiers can be used in bridges, regardless of whether bridge specifications are included. In the cases where they are not included, evaluation of the single -ended performance can be used to predict the actual bridged capability of the device.

Data Acquisition Solutions for Analog-to-Digital Converters

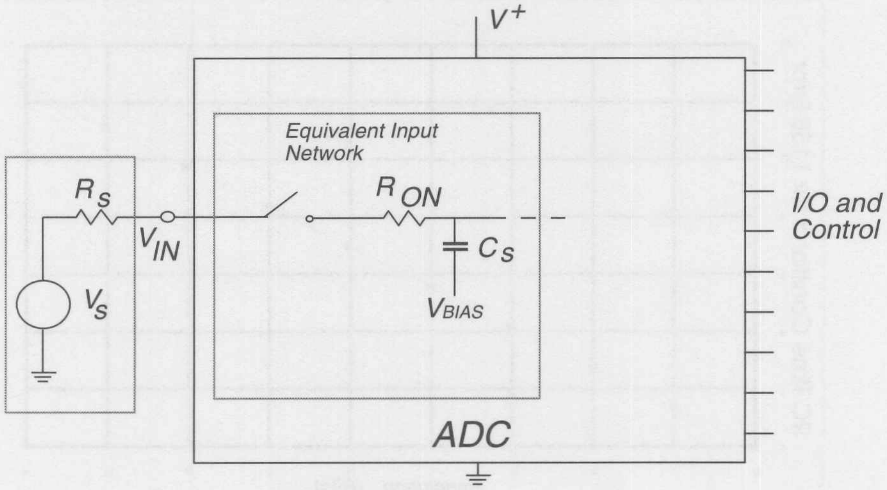
Image Scanners

Temperature Sensors

Low-Voltage ADC Applications Considerations

With the increase in demand for battery-powered electronic products, and for lower power consumption in line-powered products, analog components that operate from single 3V and 5V power supplies are becoming more widespread. To successfully design with low-voltage analog and data acquisition devices, it is important to understand how they differ from conventional components.

Low-Voltage ADC Input Stage

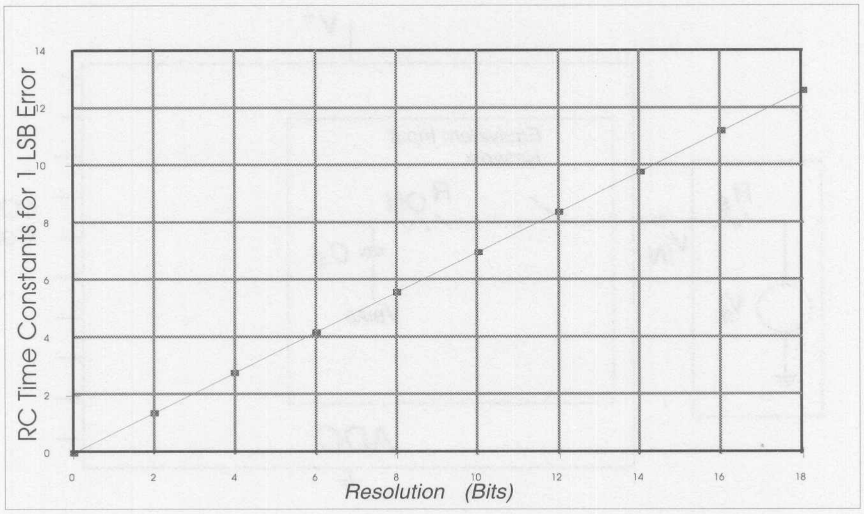


3

Nearly all ADCs designed for operation on 3V or 5V power supplies are CMOS devices, and most of those have a sampling input stage that can be represented by the network shown above. This circuit is a highly-simplified version of the real thing, but it helps illustrate some of the issues that arise when using low-voltage ADCs.

When the ADC samples an input signal, the switch closes, connecting the signal source with sampling capacitor C_S. Normally, the switch remains closed for a period of time that is controlled either by an internal timer or by an external clock or logic signal. For accurate conversions, the switch on resistance R_{ON} and the external source resistance R_S must be low enough to allow C_S to charge to within a fraction of an LSB of the input voltage during the sampling period. If the R-C time constant is too large, errors due to incomplete charging will occur.

Input Stage Errors Increase with Source Resistance



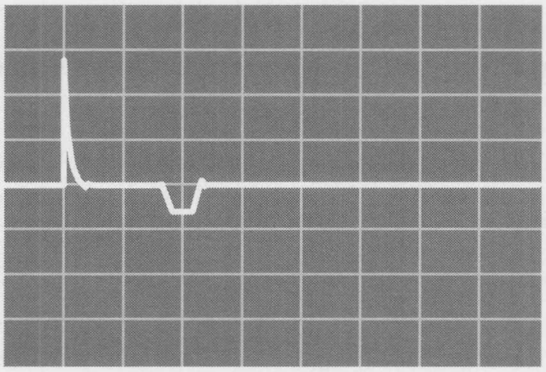
The circuitry that drives this network will directly influence system performance. As we mentioned earlier, a resistive source must charge the sampling capacitor during the ADC's sampling window. Large source resistances will result in incomplete charging, leading to conversion errors.

As the source resistance increases, so do errors resulting from incomplete charging of the sampling capacitor. The more resolution we expect from a converter, the more time we must allow to charge this capacitor. The number of time constants needed to limit input errors to 1 LSB can be found by the formula:

$$TCs = -\ln(2^{-n}),$$

where n is the converter resolution in bits.

Typical Low-Voltage Amp-ADC Combination



LMC6482 Driving ADC12038. 200mV/div, 500ns/div.

The transient nature of the ADC's input stage also affects active circuitry driving it. Very slow amplifiers, or fast amplifiers with poor settling performance, may not be appropriate for use with low-voltage ADCs. The LMC6482 is a dual low voltage CMOS amplifier with rail-to-rail input and output swing capability. When driving the ADC12038, it settles to 12 bits in about 1μs, which is compatible with acquisition time of the ADC12038. Note that two transients are visible here. The first one is due to the input mux, and occurs before the acquisition period begins. The second is due to the sampling process.

ADC Input Interface Solutions

- If amplifier or RC settling takes longer than internally-set acquisition time:
  - Slow down ADC clock (just during acquisition time if throughput rate is critical).
  - Use ADC or DAS chip with programmable acquisition time.
  - Buffer high source resistances with CMOS amplifier.

6

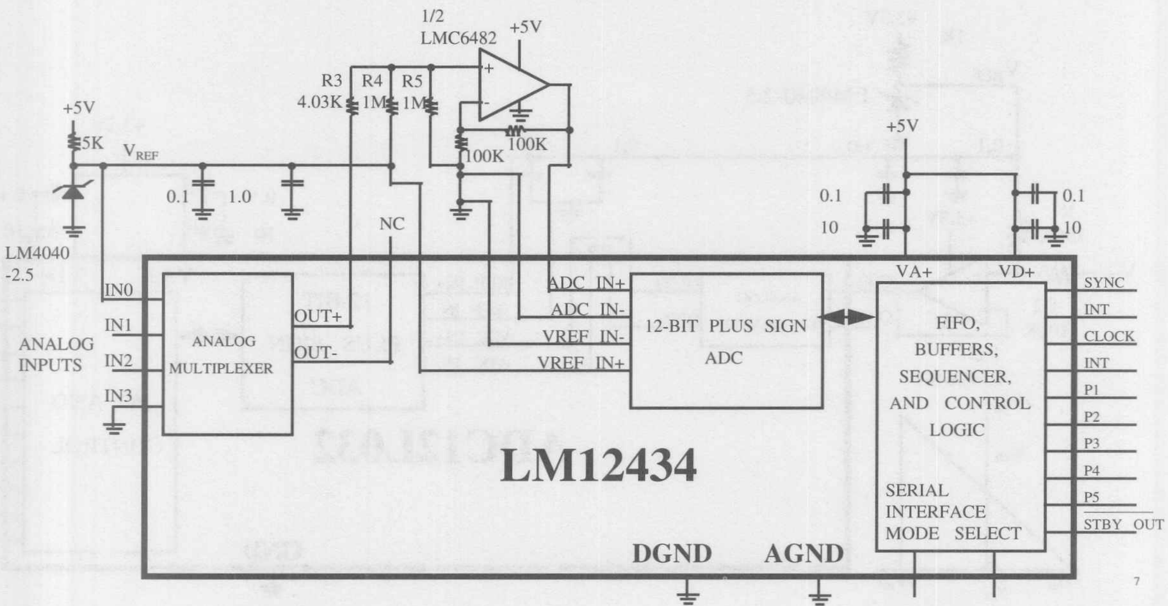
If the driving amplifier can not settle fast enough, or the input time constant is too long to allow sufficient charging of the sampling capacitor, there are three things we can do.

We can slow down the ADC clock to create a longer acquisition time. But this can slow down throughput. If throughput is critical, we may want to slow down the clock only during signal acquisition. This may not be easy to do, however.

Another alternative is to use a converter or a DAS with a programmable acquisition time. This is much easier than the first alternative.

Finally, we can buffer high source resistances with the proper CMOS amplifier.

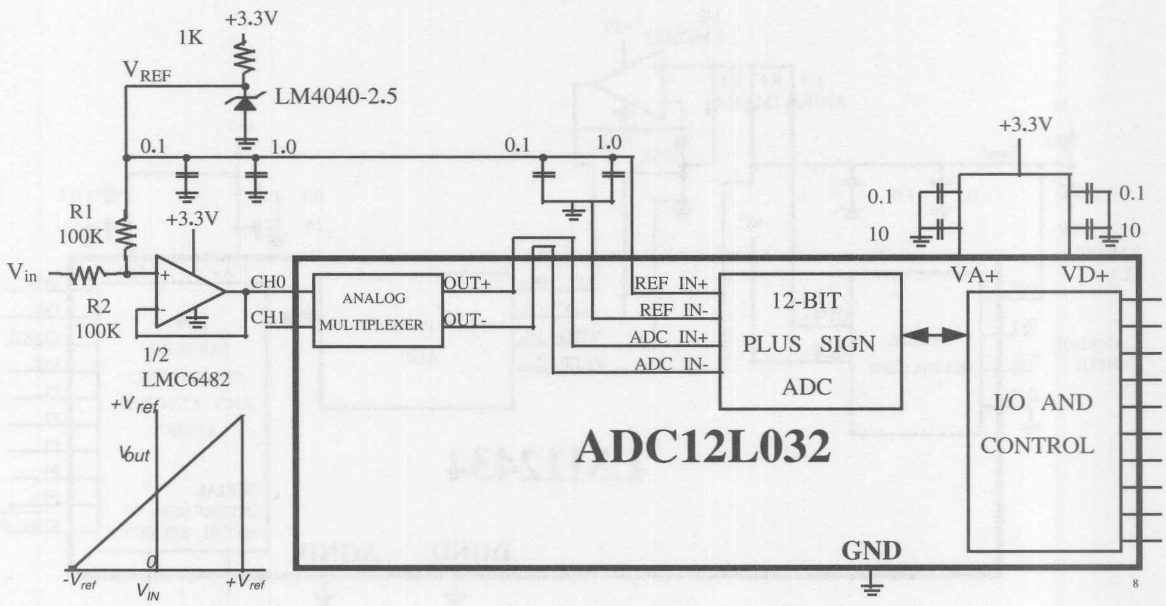
Calibrating System Errors



External analog signal processing circuitry can add to the system's overall error. In this example, a low-voltage amplifier with a few millivolts of offset error provides gain for two analog input signals, which would be connected to IN1 and IN2.

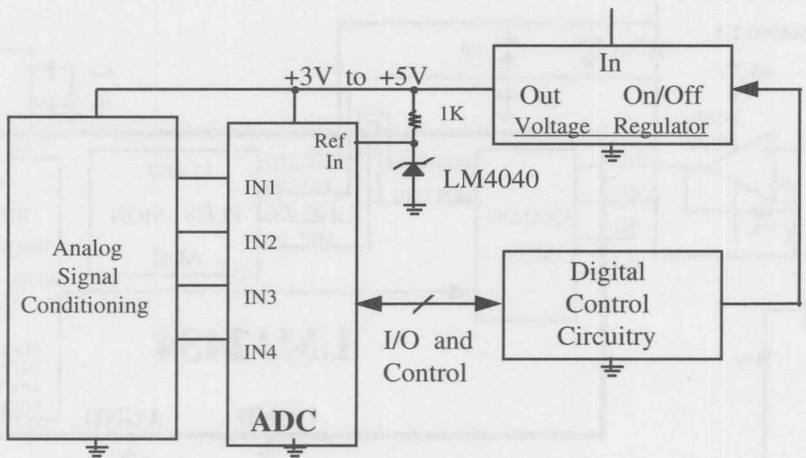
The amplifier's offset error can be measured by connecting mux IN3 to the multiplexer's OUT+ and performing a conversion. This yields a digital correction value that can be subtracted from future conversion results. Measuring the output when IN0 is connected to multiplexer's OUT+ allows us to compensate for gain errors as well.

Handling Bipolar Inputs



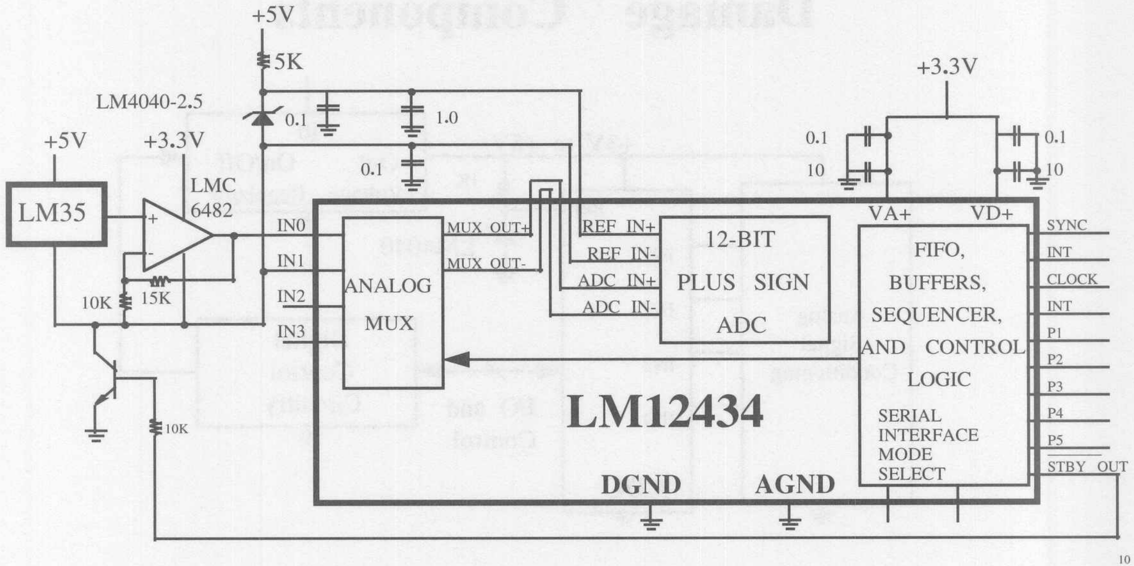
It is possible to convert bipolar signals with a converter that uses a single supply. The simplest way to do this is to offset and attenuate the input signal with a voltage divider connected to V_{REF}. The Amplifier's input voltage now swings from 0V to V_{REF} as the input voltage swings from $-V_{REF}$ to $+V_{REF}$.

Simple Shutdown Techniques Can Damage Components



This is one way to power down a circuit when it is not needed. Using this type of circuit can damage analog components that are connected to digital control circuits if those circuits remain active while the analog power supply is off.

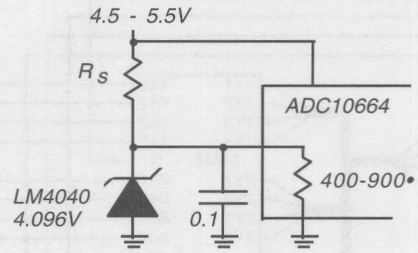
Better System Shutdown



A better method to power down circuitry that is not needed is to use a component that can be shut down, then use that device to shut down all other devices that will not be needed.

Here, the Standby Output of LM12434 serial Data Acquisition System (DAS) is used to shut down all analog input devices. The DAS, when told to go into the low power standby mode, brings its Standby Output pin low until it is returned to the active state.

Shunt references near V^+ should be lightly loaded



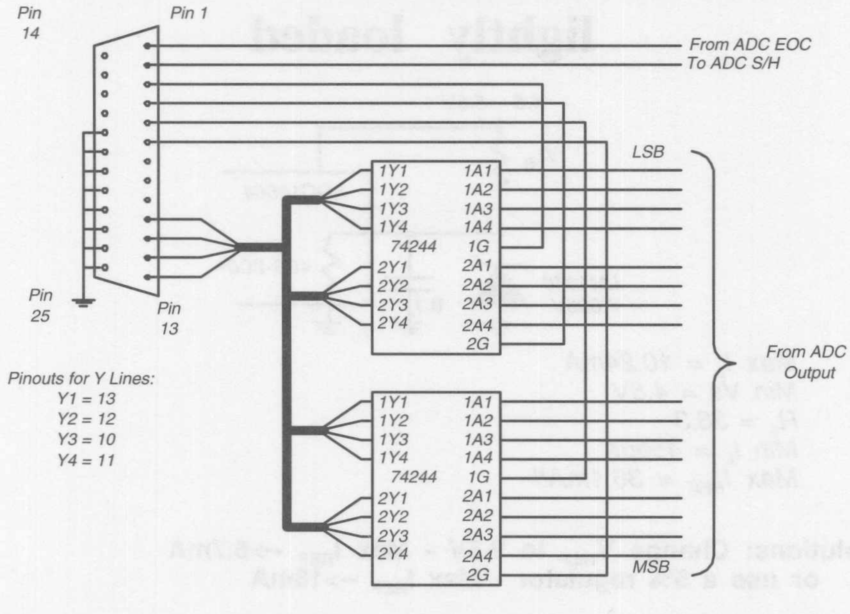
Max I_L = 10.24mA
Min Vs = 4.5V
R_s = 38.3
Min I_L = 455µA
Max I_{REF} = 36.1mA!!

Solutions: Change V_{REF} to 2.5V - Max I_{REF} -->6.7mA
or use a 5% regulator - Max I_{REF} -->18mA

Shunt voltage references are an excellent choice for low-voltage designs, because of their low input voltage headroom requirements. But when the input voltage is near the reference output voltage, be careful. If your load current and supply voltage can vary appreciably, you can end up with more current flowing through the reference than the reference can safely accomodate.

In this example, R_S is small enough to provide sufficient current when the supply voltage is at its minimum and the load current is at its maximum. When the supply voltage is at its maximum and the load current is at its minimum, however, the current that must be absorbed by the reference is excessive - in this case almost double the absolute maximum rating for the reference.

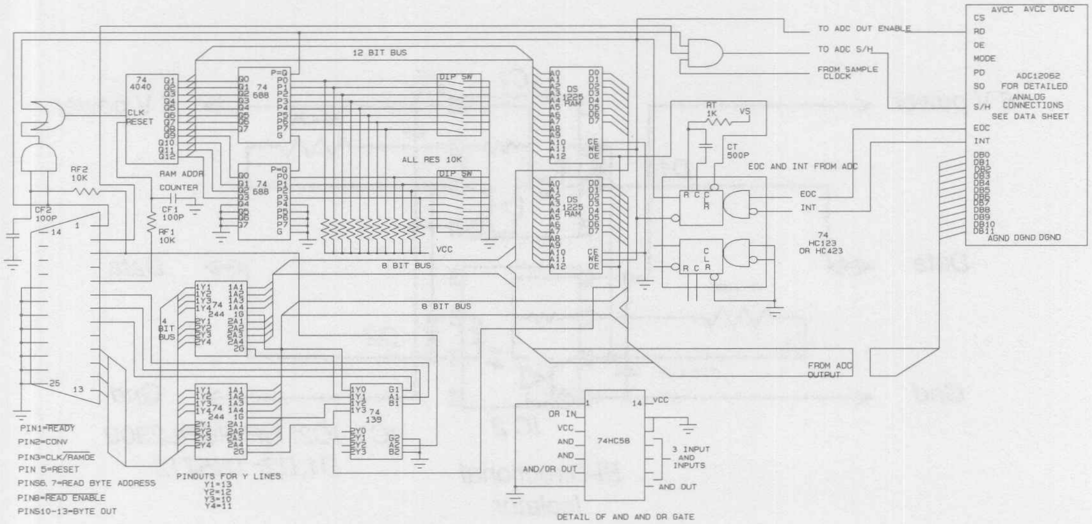
Simple ADC-to-PC Interface



SIMPLE UNIVERSAL ADC TO PARALLEL PORT INTERFACE

The parallel printer port makes a desirable, if slow, interface. This interface represents the simplest possible "universal" interface for a parallel printer port to connect to ADC's with up to 16 bits. The line to the ADC S/H control input is used to initiate conversion and the EOC line signals to the computer that the data is ready. The data is then read in four 4-bit bytes.

Interfacing Fast ADCs to PCs



13

The speed limitations of the parallel printer port can be overcome by storing some number of ADC samples in RAM at the conversion rate of the ADC, then downloading them to the parallel port at its own slower pace. This circuit allowed for a programmed storage of up to 4096 samples which it would capture automatically on command. This circuit provides an interface to a 12-bit, 1MSPS ADC12062, and downloads to the parallel port in 4-bit bytes.

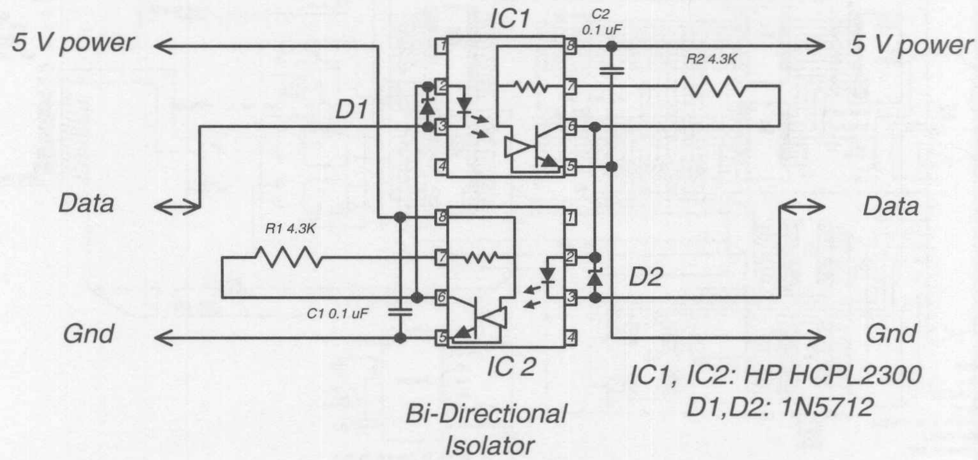
When pin 1 of the DB25 connector goes high, it enables the S/H command to the ADC, which comes from its Sample Clock, via the NAND gate in the upper right hand corner.

Every time the ADC completes a conversion, its INT line triggers the one-shot. The one-shot takes active the Write Enable of the RAM and the ADC data is written into the RAM. At the end of the Write, the 744040 counter is updated by the falling edge of the one-shot output.

When the counter has reached a value equal to the preset DIP switches (0 to 4095), the output of the 74688 comparators inhibits the ADC S/H commands.

To retrieve the data, the counter is reset. Pin 3 of the DB25 enables the RAM to output data, which is then fed to the computer in 4-bit bytes via the 74244's. After reading a complete ADC word, the computer takes Pin 3 low and this increments the counter. The computer should monitor Pin 1, when it goes low all data has been retrieved.

I^2 C® Isolation



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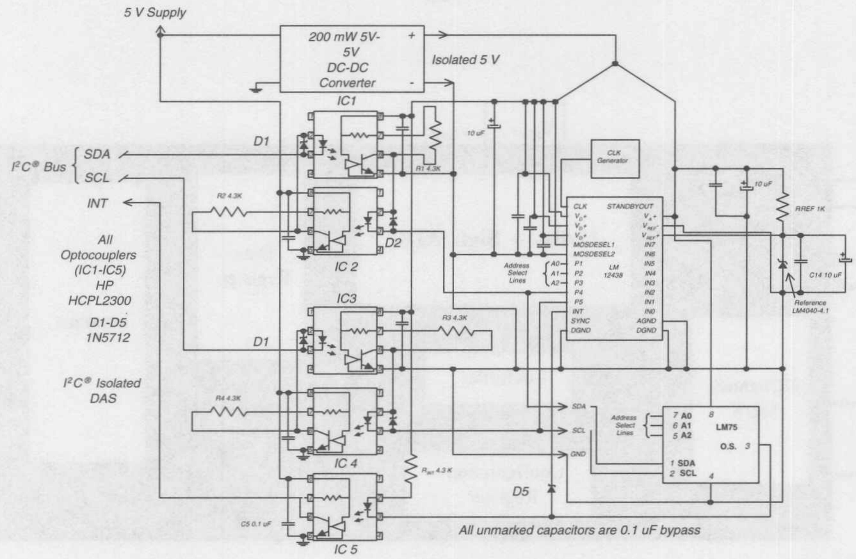
In hostile electrical environments where ground loops and safety are issues, isolation of key components can be a life saver. The I^2 C© interface is finding increased popularity and is an excellent candidate for isolation. This bus has bi-directional data and clock lines, so when it needs to be isolated, a bidirectional isolation circuit like this one is necessary. One bidirectional isolator circuit, as you'd use on the data line, is shown. A second circuit must be used on the clock line.

The I^2 C© bus uses open-collector lines, with a high rest state. If the left side data line goes low, current flows via the photodiode of IC1, activating IC1's output transistor, taking the output low. If IC1's transistor sinks current from the right side data line, this current reverse biases IC2's photodiode, and D2 is included to keep the reverse drop to less than 0.4 volts. This prevents the signal from feeding back in IC2 and latching the circuit. Taking the right side data line low causes a mirror image of this action to occur.

I^2 C© is a registered trademark of Philips Corporation

I² C® Isolated Data Acquisition

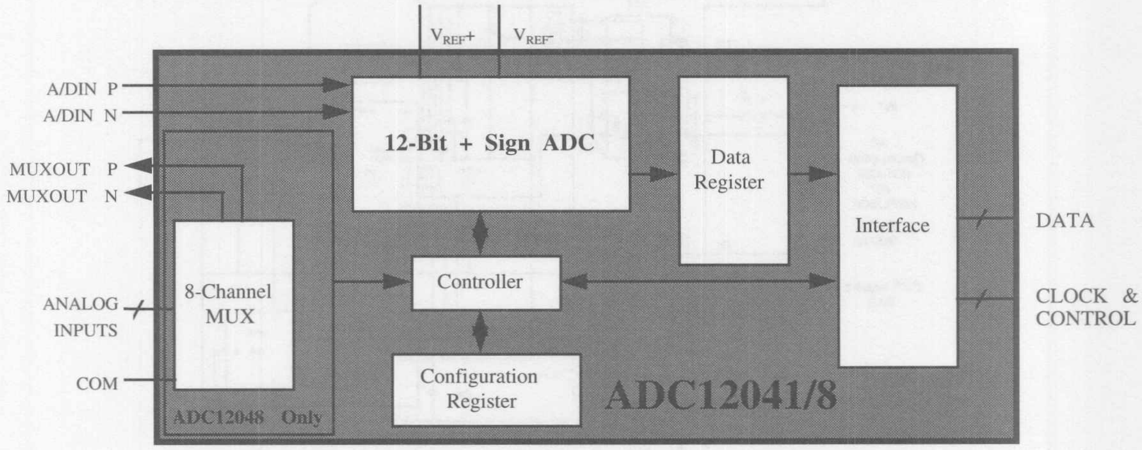


A lot is happening here. First of all, we have an isolated I² C© bus. On this bus, we have an LM12438 Data Acquisition System, or DAS. In addition, we have an LM75 temperature sensor. By doing this we have avoided using one of the analog inputs of the LM12438 to measure temperature, as the LM75 provides a ready-to-use digital word as its output.

A system such as this would be especially useful for thermocouple data acquisition. The LM75 would be mounted where the thermocouple leads terminate, and this temperature used to perform cold-junction compensation in software (note that the DAS inputs are not thermocouple-ready, considerable additional gain is needed to interface to thermocouples).

The INT of the DAS and O.S. output of the LM75 are "or'd" to the interrupt line. The diode, D5, is necessary since the DAS INT line is a totem-pole output. The O.S. output of the LM75 is a programmable overtemperature shutdown with default limits of 80 degrees for temperature with a 75 degree hysterisis point.

ADC12041/ADC12048 Block Diagram



16

The ADC12041 and ADC12048 are 12-bit-plus-sign ADCs with a parallel digital interface that allows easy, "no glue logic" interface to several popular microcontrollers and DSP chips. The ADC12048 includes an 8-channel analog multiplexer. The ADC12041 and ADC12048 feature a fast, 222kHz throughput rate and low 34mW power consumption. A shutdown mode reduces power consumption to just 75uW.

The ADC1204x has two registers, an 8bit (13bit for the ADC12048) write only **Configuration** register and a 13bit read only **Data** register.

The **Configuration** register is used to program the functionality of the ADC1204x, e.g. the MUX channel selection (ADC12048 only), the acquisition time, the bus width, the SYNC mode and the mode of operation (Full calibration, Auto-zero, Standby, Reset and Conversion). It is accessed by the microprocessor or microcontroller simply by writing to the ADC1204x.

The **Data** register holds the 13bit conversion result, which is in signed two's complement format. The **Data** register is accessed whenever the ADC1204x is read from by the mP/mC.

The WMODE pin is used to determine the active edge of the write signal (WR).

These ADCs can interface directly to the following processors with no "glue logic":

National's HPC

Intel386 and 8051

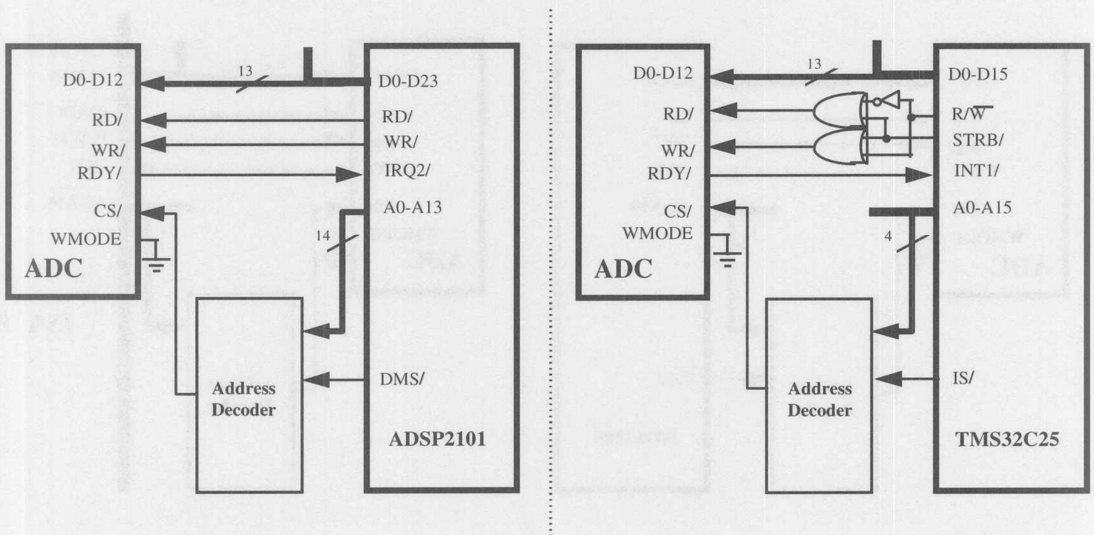
TMS320C25,

Motorola MC68HC11/16

Hitachi 64180

Analog Devices ADI121xx.

ADC12041/ADC12048 Microprocessor Interface

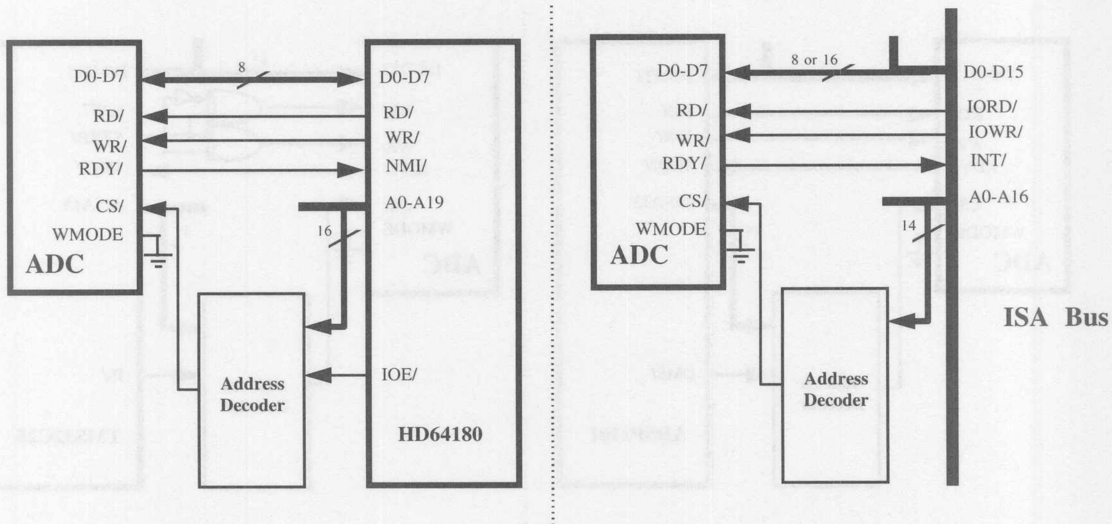


17

ADSP2101 has a 24 bit data bus. 13 of those bits can be used to interface to the ADC12041/ADC12048. DMS/ (Data Memory Select)is an active low signal that indicates that the data memory bus has been selected.

The TMS320C25 has a 16-bit address bus. Only the lower 4-bits (A0-A3) are used to address I/O space.

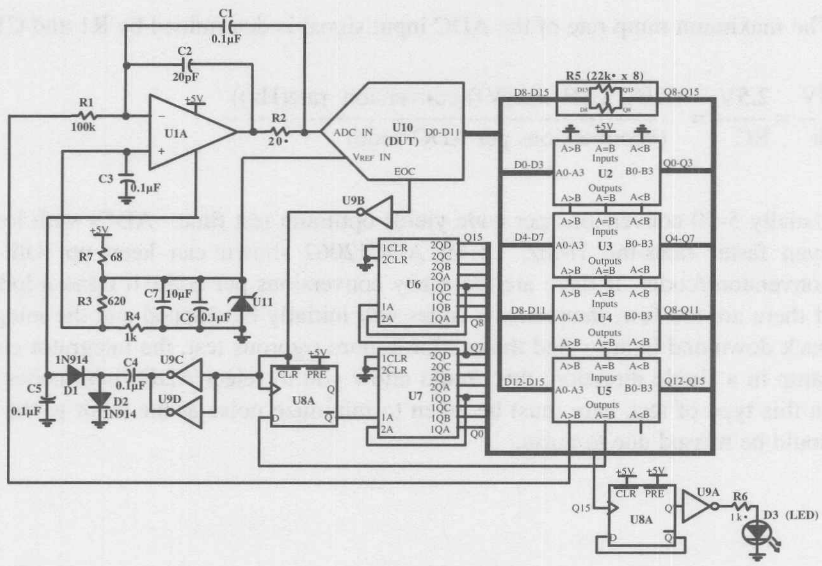
ADC12041/ADC12048 Microprocessor Interface



18

The HD64180 has a 20-bit address bus, 16 bits of which are used to address I/O space.

Simple Missing Code Test Circuit



U1	LM6142
U2-U5	74HC85
U6-U7	74HC393
U8	74HC74
U9	74HC04
U10	ADC12062
U11	LM4040-4.1

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This circuit will test any 8 to 16 bit A/D converter for missing codes. Advantages over similar designs include the ability to "hunt" up and down for codes, operation with 8 to 16 bit ADCs without any modifications, single +5V supply operation, and an inexpensive design consisting primarily of commonplace discrete logic.

The design consists of a 16 bit counter (U6 and U7), a 16 bit digital magnitude comparator (U2-U5), and an integrator (U1). The counter output is compared to the output of the ADC. If the counter is higher than the ADC output code, the comparator's A>B output is low, causing the integrator output (the ADC input) to ramp up. If the counter is lower than the ADC output code, the A>B output goes high, pushing the integrator output down. In this way the input voltage to the ADC is servoed to force a certain output code (in much the same way an ADC would be used in a control system, where no missing codes is a must). If that code is missing, the integrator will hunt up and down forever and the counter will never increment. If the ADC produces that output code, then the A=B line of U5 will go high, and the counter is incremented to the next code.

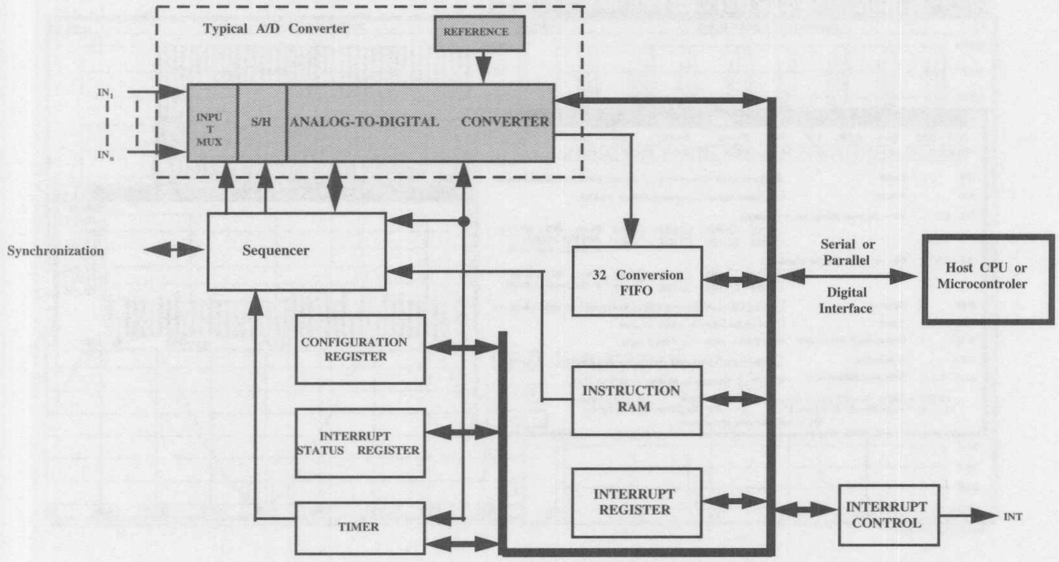
The ADC's End Of Conversion (EOC) signal is used to clock U8A and prevent any spurious signals due to state changes in the counter, ADC, or comparator logic from propagating through and erroneously incrementing the counter. The ~100ns propagation delay of the comparators limits this circuit to ADCs with conversion speeds of ~10MHz or lower. The output of *Q is the increment counter signal whose pulse width is equal to the amount of time that EOC is low. R5 insures that unused ADC bits (for ADCs with less than 16 bit resolution) are always equal to the current count. U8B changes state (flashing D3) every 65,536 counter codes, visually indicating that the ADC has no missing codes. U9C and U9D form a voltage doubler to generate a small (about 3.5V) negative supply to give the LM6142 additional drive when providing a 0V output voltage. R2 and C2 improve the performance of the op amp when driving capacitive loads typical of ADC inputs.

The maximum ramp rate of the ADC input signal is determined by R1 and C1 using the equation:

$$\frac{dV}{dt} = \frac{2.5V}{RC} = \frac{(\text{ADC LSB size(V)})(\text{conversion rate(Hz)})}{(\#\text{conversions per ADC code})}$$

Usually 5-10 conversions per code yields optimum test time. ADCs with low DNL can be tested at even faster rates-the 1Mhz, 12 bit ADC12062 shown can keep up with a ramp as steep as 2 conversions/code. If there are too many conversions per code, it takes a long time to test all codes. If there are too few, then narrow codes will initially be skipped and the integrator will have to ramp back down and hunt to find them. For a more rigorous test, the integrator could be designed to only ramp in a single direction; this would allow you to reject ADCs with codes below minimum width. In this type of test, care must be taken to minimize noise at the input to the ADC, otherwise a code could be missed due to noise.

Data Acquisition System on a Chip



A Complete Data Acquisition System

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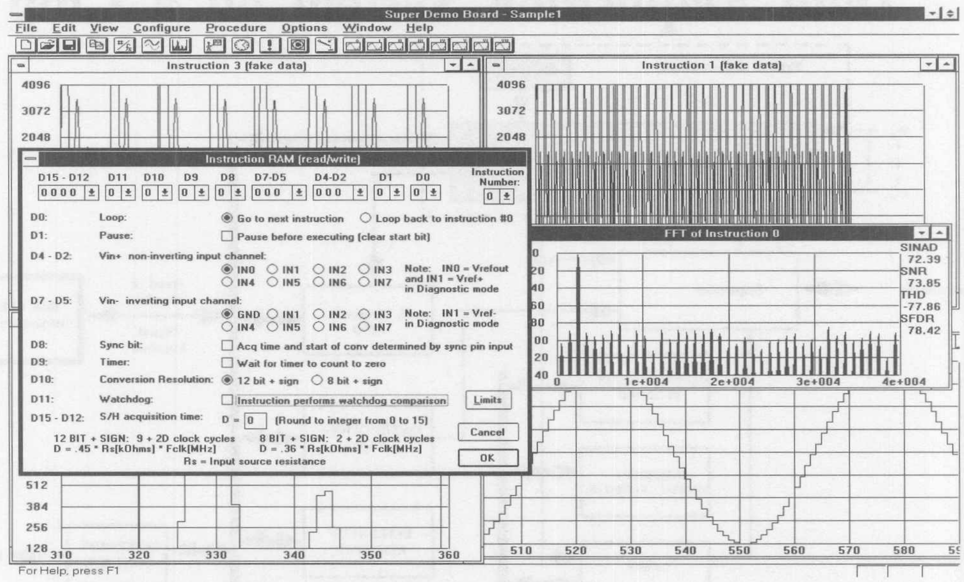
What is a DAS? A DAS is a Data Acquisition System on a chip. Above is a block diagram for National's LM12458 and LM12438 DAS families. These devices are true Data Acquisition Systems. They acquire and store data from up to eight analog sources, interrupt the host processor upon occurrence of pre-programmed events, and even perform watchdog comparisons, looking for out of limit conditions and low power supply voltages.

The Sequencer, or Controller, is a state machine that takes instructions from the Instruction RAM and acts accordingly. The contents of the Configuration Register, Instruction RAM, and the Interrupt Enable Register may be programmed to determine what signals to compare or convert, and when to report back to the host processor with an interrupt signal. Once the DAS has been started, no further "baby sitting" is required by the host CPU until the occurrence of specifically programmed events.

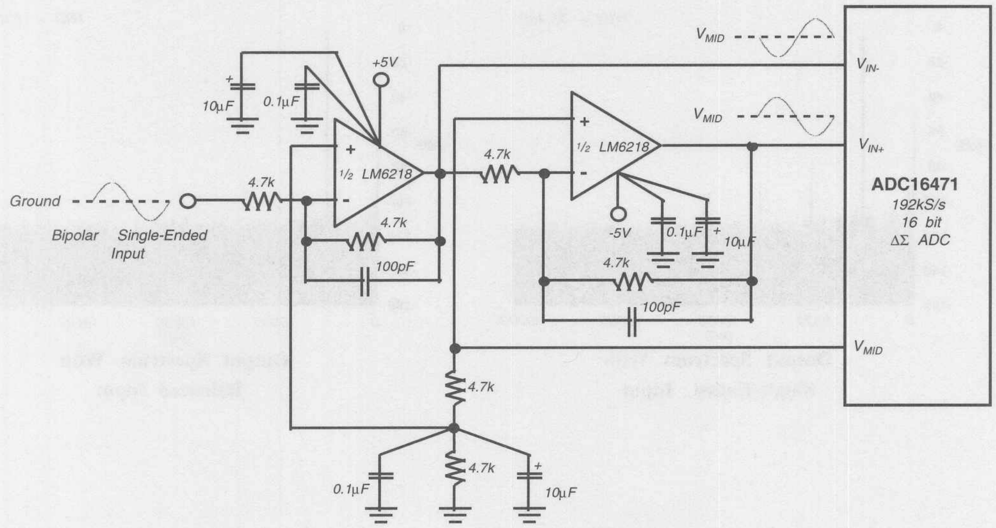
WaveVision Evaluation Software



For help in the evaluating the DAS, National provides an evaluation board, and the freely available WaveVision software. These tools quickly give an intuitive feel for the operations of the DAS products and serve as reference designs. The board/software combination allows you to program the registers of the DAS , acquire analog conversion data, and evaluate the results via FFT and dynamic performance calculations within the Windows environment. The evaluation board easily connects to the PC through a serial communications port.

Similar evaluation tools are available for the ADC12048, ADC12038, and the LM75.

Single Ended to Balanced Buffer



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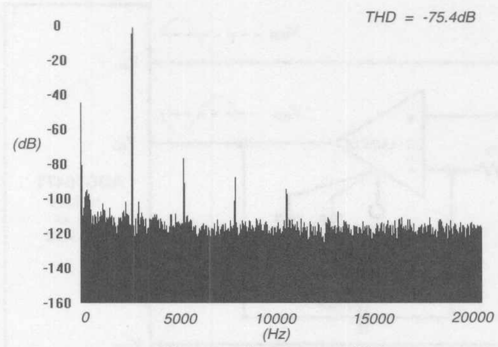
Differential input ADCs exhibit the best distortion performance when balanced differential signals are applied to their analog inputs. When ADCs are driven by differential signals, the conversion process will cancel out common mode noise and reduce harmonic distortion.

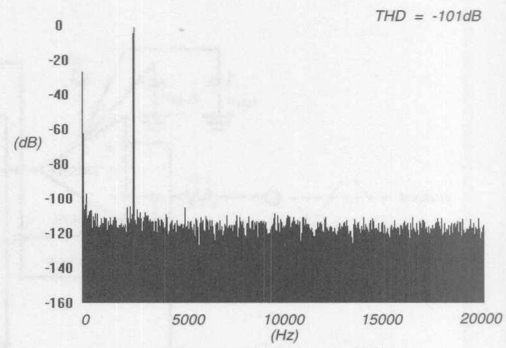
This circuit can be used to convert a single-ended signal centered around ground to a balanced signal centered about the middle of the ADCs analog input range. The circuit is shown driving an ADC16471, which is a 16 bit, 192kS/s, ΔΣ ADC. In this case, the output signals, at the V_{IN+} and V_{IN-} pins of the ADC16471, will be centered around the ADC16471's half supply output voltage, V_{MID}.

For input signals that are not centered around ground, capacitor-couple the input signal.

Single-Ended vs. Balanced Input



Output Spectrum With Single-Ended Input



Output Spectrum With Balanced Input

24

Driving the ADC16471 with a balanced differential input signal, as with any differential input ADC, will cause a cancellation of common mode noise and reduce harmonic distortion. Here we see the circuit of the previous page yields a distortion performance (THD) improvement of 25dB compared to a single-ended input signal.

Introducing the ADC16071/ADC16471 ΔΣ 16-bit 192ks/s ADCs

- THD:
  - (48kHz output data rate) -94 dB (typ)
  - (192kHz output data rate) -80 dB (typ)
- Digital Filter:
  - Passband Ripple 0.005 dB (typ)
  - Stopband Rejection 90 dB (typ)
- Power Consumption:
  - (48kHz output data rate) 275 mW (max)
  - (192kHz output data rate) 500 mW (max)
  - Power down 6.5 mW (max)
- Output Data Rate 7 kHz to 192 kHz
- Serial Data Interface
- Single 5V supply

National's ADC16071 and ADC16471 are the world's fastest ΔΣ 16 bit analog to digital converters with an output data rate of 192KHz (min). They feature a 4^{th} order modulator with 64 times oversampling.

The ADC16471 has a built-in fully differential bandgap voltage reference with a fixed analog input range of +/-2.5V. The ADC16071 requires an external reference with a user selectable analog input range.

The ADC16071/ADC16471 have a serial digital output that requires zero glue logic to interface to popular DSPs including Motorola's DSP56001, TI's TMS320C3x and Analog Devices' ADSP-2101.

Two ADC16071/ADC16471s can easily be configured to share a single serial data line and operate in a "stereo", or two channel multiplexed mode.

CLC949 A/D Converter
12 bits, 20 MegaSamplesPerSecond

- **Lowest Power High Speed Converter Available**
- **Single Supply Operation**
- **Excellent Dynamic Specifications**
  - **– 72dB SFDR**
  - **– 68dB SNR**

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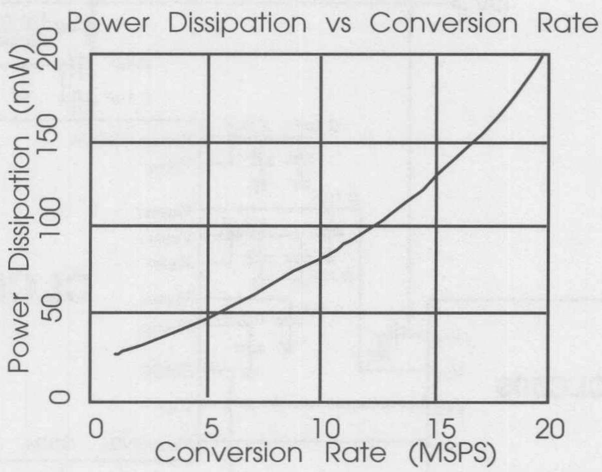
The CLC949 is a high performance, 30MHz, 12-Bit A/D converter sporting excellent dynamic specifications and low power. Control of the exact power consumption is left with the user through the two "BC" input settings and resistor R_P in series with the BIASC line, as well as through selection of the conversion (clock) rate.

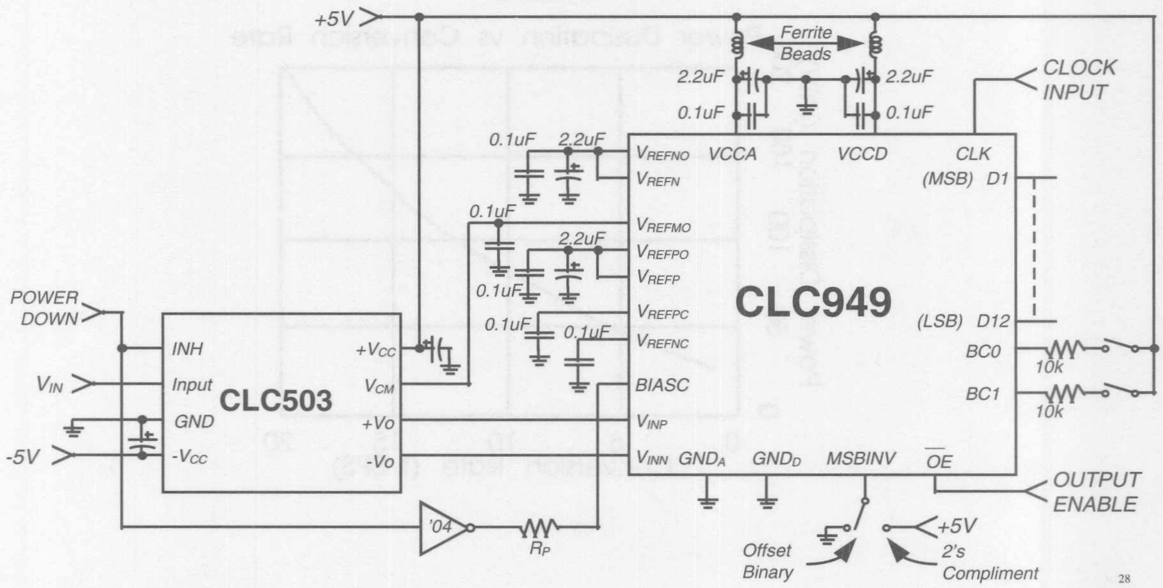
The CLC949 offers power consumption that is only 20% that of any other high-speed 12-bit ADC operating at the same speed. In addition, while most other converters in this speed and resolution class require dual power supplies, the CLC949 operates from a single 5V power supply.

Power Depends on Conversion Rate



The CLC949's power consumption depends upon the conversion rate. By adjusting its operating current to match the conversion rate, you can achieve extremely efficient conversions over a wide range of speeds.

CLC949 With CLC503



The CLC503 simplifies the analog interface to the CLC949 by providing a single chip (8 pin SOIC) solution to the analog front end. The CLC503 provides buffering and level-shifting for ground-referred input signals with no degradation of system performance. The CLC503 is perfectly suited to the task of single-ended to differential signal conversion for the CLC949. Its wide bandwidth (100MHz), low distortion (-80dBc @10MHz, $4V_{P\text{-}P}$), and fast settling (15ns to 0.01%) means gaining the performance advantage of differential inputs to the CLC949 while retaining the CLC949's bandwidth and linearity.

Both of these devices can be powered down when they are not needed.

CLC532/533 Buffered High-Speed Analog Multiplexers (2:1, 4:1)

- Very fast settling time (17ns to 0.01%)
- Very high Channel-Channel Isolation (80dB @10 MHz)
- Low Distortion (-80dB @ 5 MHz)
- High Bandwidth
  - 180MHz Small Signal
  - 45MHz Large Signal
- Low Noise (-42 μVrms)
- More than just a mux!

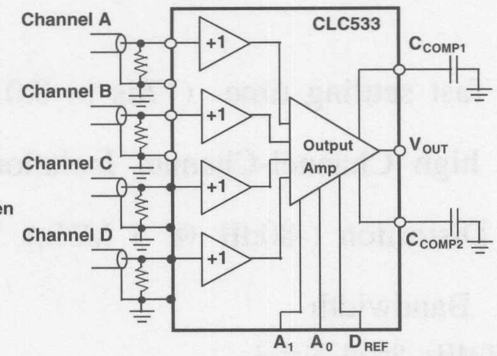
29

These high-speed, low-distortion, low-noise buffered multiplexers provide the ultimate in performance for fast analog and data acquisition applications.

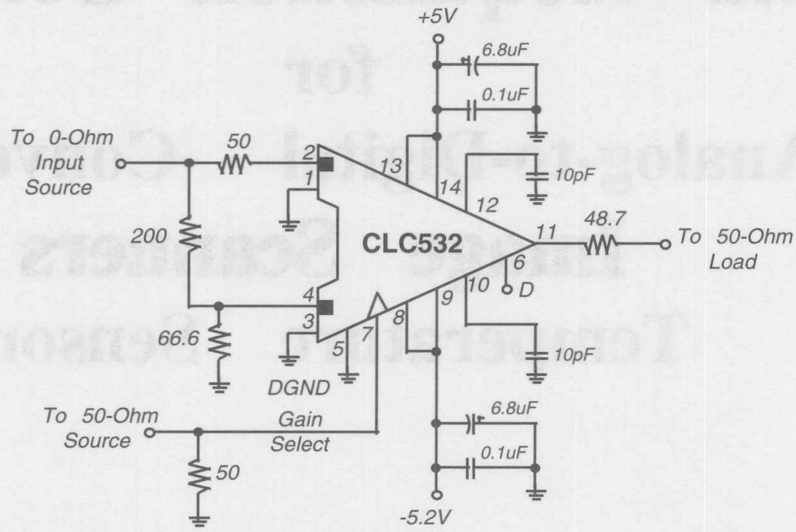
CLC533 Functional Diagram

A_1	A_0	OUT
0	0	A
0	1	B
1	0	C
1	1	D

ECL Mode - D_{REF} = open
TTL Mode - D_{REF} = +5V



Selectable Gain Stage Improves ADC Dynamic Range



In many applications, such as RADAR, the dynamic range requirements may exceed the accuracy requirements. Since wide dynamic range ADCs are also typically highly accurate ADCs, this often leads the designer into selecting an ADC which is a technical overkill and a budget buster. By using the CLC532 as a selectable-gain stage, a less expensive ADC can be used. As an example, if an application calls for 80dB of dynamic range and 0.05% accuracy, rather than using a 14-bit converter, a 12- bit converter combined with the circuit in figure 8 will meet the same objective. The CLC532 is used to select between the analog input signal and a version of the input signal attenuated by 12dB. This circuit affords 14-bit dynamic range, 12-bit accuracy and 12-bit ease of implementation.

Data Acquisition Solutions
for
Analog-to-Digital Converters
Image Scanners
Temperature Sensors

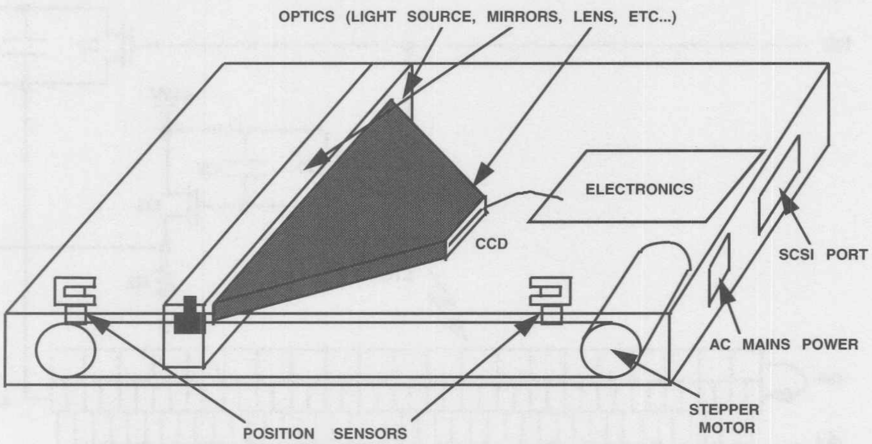
32

Why are we talking about CCD document scanners? Because, thanks to the ubiquity of computers everywhere you turn and the popularity of the World Wide Web, the need for converting printed images to digital format has never been greater.

National Semiconductor has spent a great deal of time learning about this market and has introduced an integrated CCD processor that gives scanner manufacturers (and their customers) higher performance for lower price.

While you may not be directly involved in the scanner industry, we hope you'll find this discussion an interesting example of National Semiconductor's capabilities in mixed-signal integration for signal processing applications.

CCD Document Scanners?



33

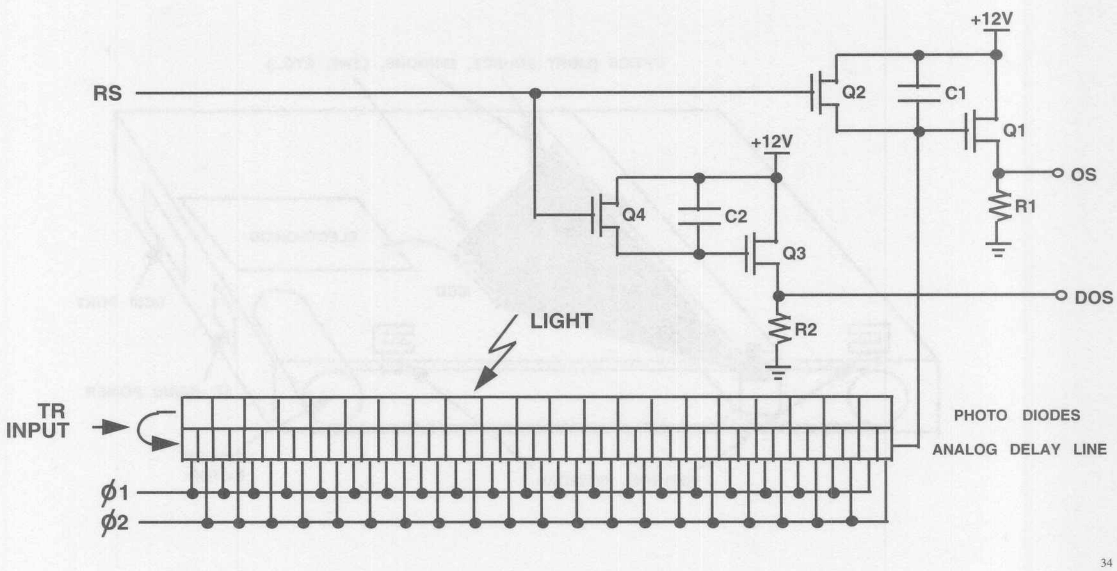
Document scanners typically consist of a linear (one dimensional) CCD image sensor which is either moved across a two dimensional image (as in handheld and flatbed scanners), or motionless while the image is moved across the CCD (as in sheetfed scanners).

The image sensor is made up of a row of photodiodes alongside a shift register used to clock the signal from each pixel out through a single output pin. The number of pixels in the sensor is determined by the width of the area to be scanned and the desired resolution. A letter-sized flatbed scanner scanning at 300 dpi (dots per inch) would require a minimum of 2550 (300pixel/inch x 8.5inch) pixels. Resolution in the other dimension, the direction of movement, is largely determined by the speed at which the sensor moves over the page.

Keeping the sensor one dimensional (as opposed to two) has many advantages, not the least of which is low cost-it is much cheaper to manufacture a 2550 pixel sensor than a 2550 x 11inch x 300pixel/inch = 8.4million pixel sensor!

These CCDs and their related optics are not inherently perfect - they require a good deal of analog and digital processing before the image seen on your computer screen looks like the original image you just scanned in. This presentation explains how the raw CCD output signal is converted to a high quality, error-corrected digital output that closely reflects the original image.

CCD Basics



Most image scanners available today use a CCD (Charge-Coupled Device) as the sensor for turning light (photons) into an electrical voltage.

Here's a simple representation of a CCD and its output stage. During the integration time, the light is turned into electrons by a photodiode array-one photodiode for every pixel. The number of electrons in each potential well at the end of the integration time is proportional to the average amount of light falling on that pixel during the integration time, hence the term *integration* time. This is useful because it allows us to determine what the average light was over a given area (defined by the width of a pixel in the direction of the linear sensor and the distance traveled by the sensor in the other).

At the end of the integration time (defined by signal TR, or TRANSFER) the electrons are transferred from the photodiodes' capacitance to the analog delay line. This is a linear array of potential wells, clocked by Ø1 and Ø2, a two phase clock. Since the electrons will follow the higher potential, the alternating clocks lead the electrons (in lemming fashion) to the output stage.

The electrons are shifted serially out of the CCD. When an electron packet reaches the end of the shift register, the electrons flow into C1, converting charge to voltage. The source follower consisting of Q1 and R1 converts this high impedance signal to a low (~1k•) impedance output, isolating the signal on C1 from the external output load. Q2 resets the voltage across C1 just prior to the next pixel's signal flowing into C1. This is controlled by the RS, or RESET, pulse.

An identical output stage, but without any shift register input signal, is used to generate a Dummy Output Signal (DOS), who's purpose will be explained later.

National Semiconductor — *Data Acquisition Applications*

Error Sources in Linear CCD Imaging

- *Light source intensity variation across CCD*
- *CCD pixel to pixel mismatch*
- *CCD Black Level Body Drift*
- *Pixel to pixel Vos*

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The signal coming from a CCD has many sources of gain error. The illumination across the page may not be even, and lenses have more attenuation on their edges than in the middle. This usually results in the digitized image being lighter in the center than on either side.

Also, the efficiency of all the photodiodes in the linear array is not identical, so for the same light stimulus two adjacent pixels could have output voltages differing by ±20%. This error manifests itself as vertical streaks in the image.

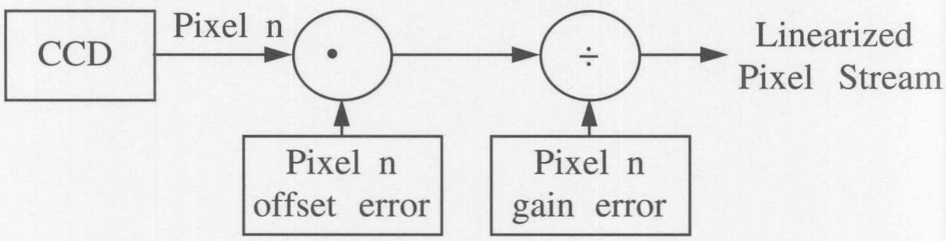
There are also offset errors to contend with. The DOS-OS subtraction does not remove all the DC offset on the OS signal, since the OS and DOS stages are not completely identical (the OS offset voltage may be signal-dependent while the DOS voltage is not). This is called black level body drift.

Many popular CCDs employ an even/odd shift register scheme, where the odd numbered pixels go through one shift register and the even numbered pixels go through another, providing an effective increase in readout speed. When the signals are multiplexed back into a single stream at the output stage there is often a different DC offset for even and odd pixels.

These offset and gain errors show up as vertical streaks or shading errors on scanned images, and much of a good scanner's design is spent on minimizing these sources of error.

Error Correction

- *Most scanning devices do some sort of calibration to correct for these errors. Higher quality scanners correct for errors on a pixel by pixel basis.*
- *The amount and type of error correction required is determined by the application and the system cost target.*



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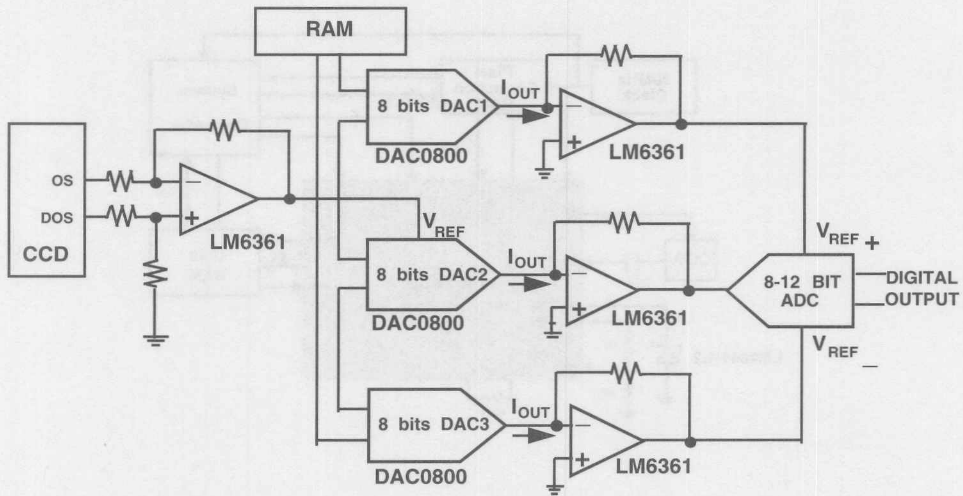
Most scanners calculate their errors by scanning a black reference and a white reference image. This image may be stripes under the glass of a flatbed scanner or a calibration sheet for use with sheetfed or handheld scanners. The variation in the output of the ADC for the black image is offset error, and the variation in the white image is gain (sometimes called *shading*) error.

Since this process gives us the gain and offset error of each pixel, we can correct distorted images scanned in the future by subtracting each pixel's offset error from the pixel stream and dividing each pixel by its gain error.

On most flatbed and sheetfed scanners this pixel-by-pixel gain and offset correction is done in hardware, with DACs doing the subtracting and multiplying.

In less expensive systems, such as hand scanners, this correction may be done digitally in the host computer's environment (this method produces a lower quality image). In some systems there may be no error correction at all.

Capturing the CCD Data: The Traditional Way



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In most current scanners, the DOS-OS subtraction is done in the first stage, providing a ground referenced, single-ended signal for the rest of the circuitry. That circuitry usually consists of gain correction and offset correction.

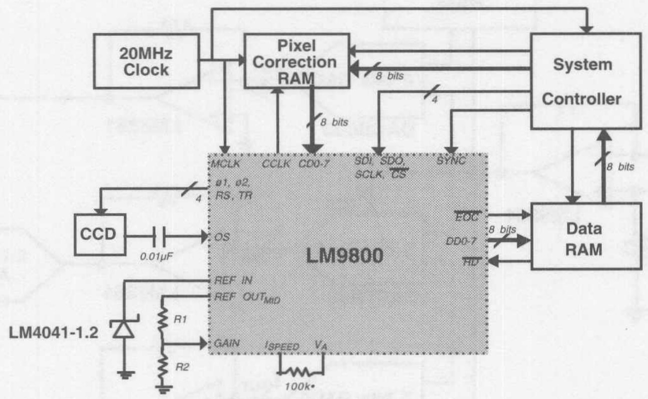
The gain correction occurs in one or two places. DAC1 modulates the ADC's positive reference voltage, effectively controlling the gain of the ADC (and therefore the effective gain on the signal). As the positive reference voltage decreases, the digital output code for a fixed DC input voltage increases.

DAC2 is used in a multiplying mode, where the DAC's reference voltage is the input signal, and the DAC code controls what portion of that input signal will pass through it. Different architectures use one or the other of these techniques; many use both. In some cases one DAC (DAC2, for example) may not change for an entire line, while the other (DAC1) changes at the pixel rate. In that case DAC2 is used as a fine gain adjust for the entire signal.

DAC3 modulates the lower reference voltage of the ADC, providing correction for CCD offset errors.

Note that both errors, offset and gain, are corrected at the pixel rate - a challenging design exercise in terms of settling times for the DACs, the ADC's reference voltages, the ADC's acquisition time, and of course the signal being digitized.

LM9800 System Block Diagram



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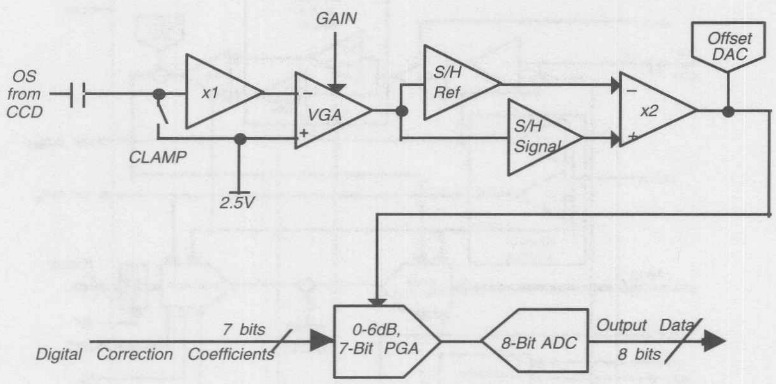
This block diagram shows a scanner system using the LM9800. Note that the only components between the CCD and the digital databus are a 0.01µF capacitor and the LM9800! The LM9800 does all the analog preprocessing, gain and offset correction, and digitization required for a high quality 8 bit image system.

The LM9800 also directly drives the clock inputs of most linear CCDs. The configuration register inside the LM9800 not only allows allows the system designer to use a wide variety of CCD sensors, it supports experimentation and fine tuning of the timing to optimize the performance of any given CCD.

On the digital interface, the coefficient bus (for the gain correction data) can be separate from or tied to the output databus for system flexibility. The EOC and CCLK signals facilitate DMA to/from external RAM.

The LM9800 is programmed through a 4-signal MicroWire/SPI serial interface to ease interfacing to popular microcontrollers.

LM9800 Block Diagram



The output of the CCD is capacitively coupled to the high impedance input buffer of the LM9800. During the optical black portion of the CCD output at the beginning of every line, the clamp is turned on, clamping the input to the buffer to 2.5V for a black input.

The VGA is used to add gain to the entire signal. This can be used to compensate for a weak light source or, in the case of a color system, the output amplitude difference between a blue line and a red or green line. This gain is fixed for a line of data; it does not vary every pixel.

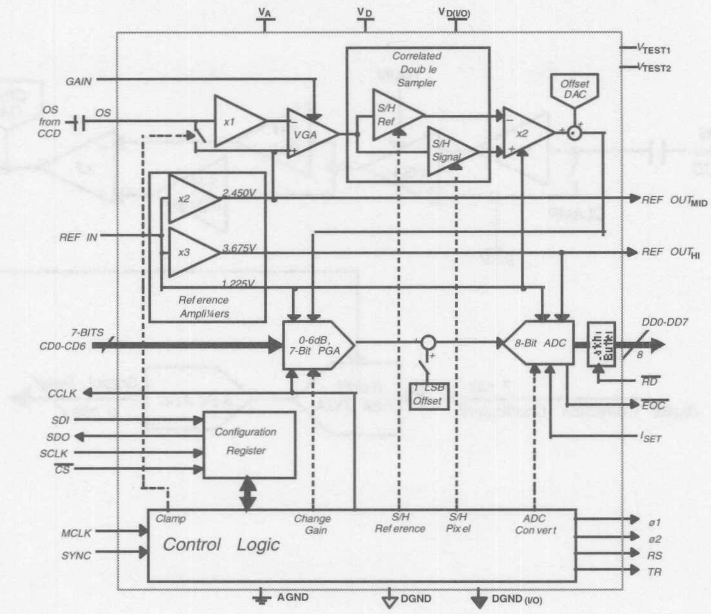
The next stage is the CDS section, shown here as two S/H amplifiers. Here the reference (black) level is sampled, the signal level is sampled, and the difference is taken and multiplied by 2 (for greater SNR in the rest of the analog chain). CDS(Correlated Double Sampling) is a higher performance technique for sampling CCD outputs than the OS-DOS method described previously.

Because there may be a DC offset error at the output of the CDS stage, a programmable DAC is added to the CDS output to null it out. Remember that CDS removes all the CCD-related offset errors, so pixel-to-pixel offset correction is not required.

The next stage is the PGA (Programmable Gain Amplifier) where the pixel-to-pixel gain (shading) correction is done. The gain is controlled by the data on the 7 bit correction coefficient databus.

The final stage is the 8 bit, 2.5MHz ADC.

The LM9800 In Detail



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This is a closer look at the LM9800.

Just about everything is configurable within the LM9800.

The CCD timing (which determines the durations of the reference level, the signal level, and the reset feedthrough pulse) is fully programmable, with 25ns precision. The output signals can also be inverted through software to support different CCDs.

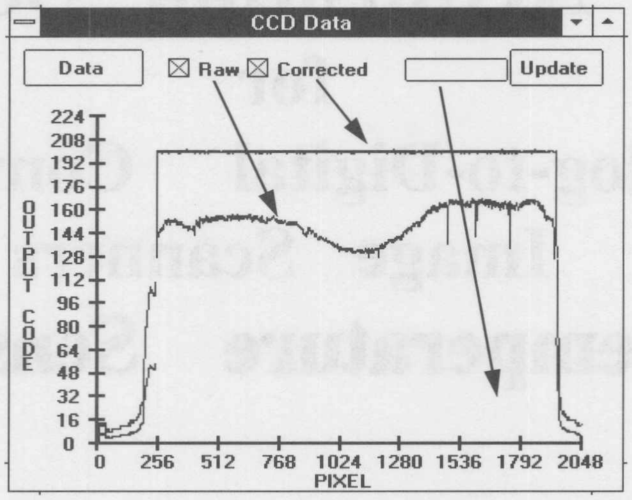
The position of the reference and signal sample pulses in the CDS section are programmable to within 25ns.

The number of dummy and optical black pixels are programmable, and there is no upper limit on the number of active CCD pixels.

There is an internal coefficient data mode to simplify calibration and system debugging during development.

All this flexibility means that the LM9800 works well not only with the sensors it was designed to work with, but other CCD and even CIS (Contact Image Sensor) sensors that were unknown to National Semiconductor during the LM9800's development.

Pixel-Pixel Gain Correction



The red (middle) trace shows the raw data coming out of a CCD sensor scanning a single white line. Ideally this line would be flat, but in practice it isn't. Here the smooth curving is caused primarily by variation in light intensity across the page, and the spikes by dust or some other opaque object in the optical path.

The green line (bottom) shows the gain correction coefficients required to linearize the CCD's output data. You can see how they are an inversion of the CCD data.

When the raw output data is multiplied by the correction coefficients inside the LM9800, the result is the corrected output shown in blue (top).

Once the input stream is normalized for the correct gain with a white input, the system will respond linearly along the gray scale between black to white.

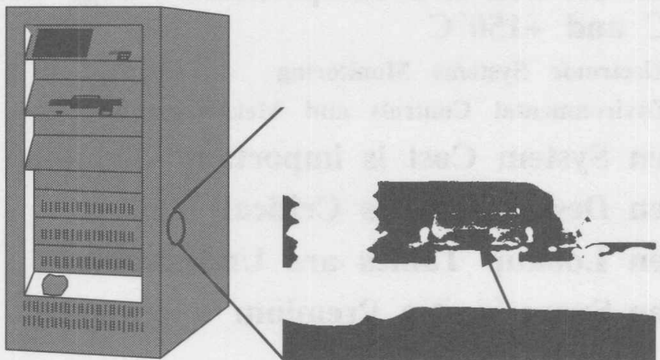
Color CCDs, with separate red, green, and blue signals, are linearized in the same way, with a white reference sheet and normalization to full scale for each color.

Data Acquisition Solutions

for

Analog-to-Digital Converters
Image Scanners
Temperature Sensors

Temperature Sensing for Improved Performance and Reliability



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In addition to conventional temperature measurement applications, temperature sensors are widely used to help prevent damage to systems due to unsafe operating temperatures. In this section we will discuss a number of sensor issues, with special emphasis on system temperature management concerns.

When Should You Use IC Temperature Sensors?

- **When the Sensor's Temperature Range will be between -55°C and +150°C**
  - **Electronic Systems Monitoring**
  - **Environmental Controls and Measurements**
- **When System Cost is important**
- **When Design Time is Critical**
- **When Lookup Tables are Undesirable**
- **When Space is at a Premium**

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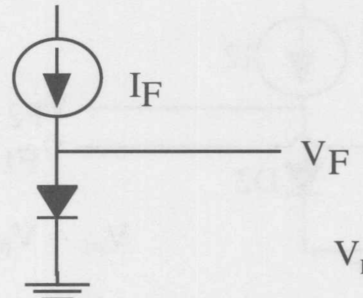
Designers have numerous options for temperature sensing techniques. Thermistors, RTDs, thermocouples, and active silicon sensors are among the most common, and each has its own set of advantages and disadvantages in any application. IC sensors have major advantages when the temperatures to be measured fall within the normal operating temperature range of silicon ICs. Among these advantages are low system cost, small size, and fast design time (because external signal conditioning circuitry is either minimal or not required). In addition, sensor ICs can include extensive additional functions, such as built-in trip-point comparators or digital I/O. And, since they include on-chip linearity correction when needed, there is no need for lookup tables to correct linearity errors.

A Very Simple Temperature Sensor



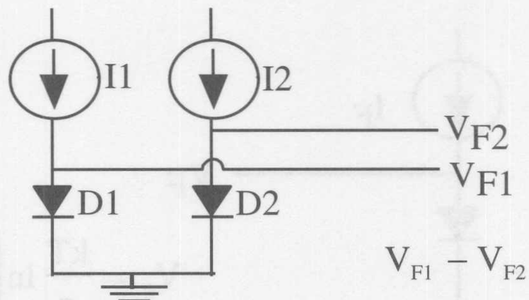
$$V_F = \frac{kT}{q} \ln\left(\frac{I_F}{I_S}\right)$$

- **Cheap and small**
- **BUT - Unpredictable "offset voltage"**

45

The rudimentary sensor shown here is simply a silicon diode driven by a current source. The diode's forward voltage drop, V_F, depends on forward current I_F, reverse saturation current I_S and junction temperature T. V_F has a -2mV/°C slope. This temperature coefficient, while much greater than that of a thermocouple or an RTD sensor, is still rather small. It is also negative, which is inconvenient in many measurement applications. Even worse, the PN junction's forward voltage has a large "offset" that varies significantly from unit to unit (because of normal process variations in I_S), so it's impossible to know what the actual temperature is without calibration at a known temperature. Nonlinearity of the temperature-to-forward-voltage transfer function is on the order of a couple of percent over the normal silicon operating temperature range.

A Better Approach



$$V_{F1} - V_{F2} = \frac{kT}{q} \ln \frac{J1}{J2}$$

Output voltage is proportional to absolute temperature

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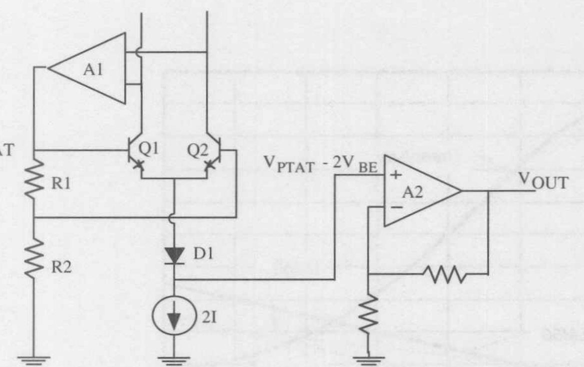
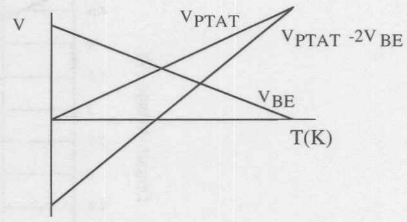
A more practical approach is to measure the difference in the forward voltages of two junctions operating at different current densities, as shown here. Different current densities can be achieved with various combinations of diode areas and currents. A typical approach might be to give D1 and D2 equal collector currents, but different areas (for example, D2 might be ten times the size of D1). The difference between their forward voltages will then be proportional to the log of the current density ratio and to absolute temperature, as can be seen in the equation above. J1 and J2 are the current densities (current per unit area) in D1 and D2 .

The differential output voltage is a fairly small voltage (typically only a fraction of a millivolt/K), so it is amplified to create a more convenient temperature coefficient, such as 10mV/K) at the IC's output. IC designers have used this basic technique for many years, beginning with the old LX5600 and LM3911, to build silicon temperature sensors with output voltages or currents proportional to absolute temperature.

A Celsius Temperature Sensor





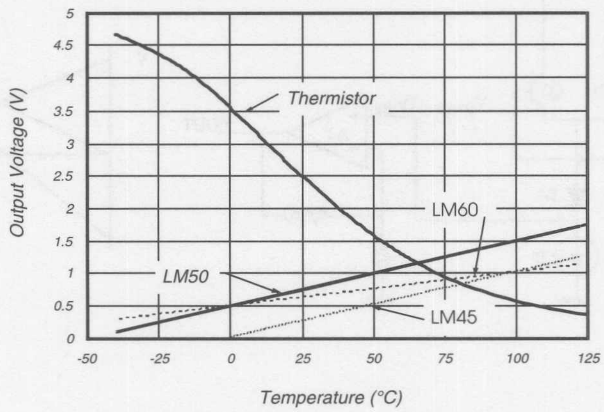
Sometimes you don't want a sensor output that's proportional to absolute temperature. Often, a measurement of Celsius or Fahrenheit temperature makes more sense in a given application. A temperature sensor IC can produce outputs proportional to one of these scales, as well.

A simplified conceptual schematic of a basic Celsius silicon temperature sensor is shown above. The sense transistors, Q1 and Q2, are scaled with a 10:1 area ratio, and current I flows through each. Amplifier A1 drives the top of R1 to force the currents through Q1 and Q2 to be equal. At 25°C, the difference in current densities causes a 60mV difference between the base voltages of Q1 and Q2 (kT/Q x ln10 = 60mV). Note that the exact value of I is not critical. As long as the ratios of the transistor areas are well-controlled, and the currents flowing through them are matched (both achievable with conventional IC manufacturing processes), the correct voltage differential will be present. This voltage appears across R1. Since 60mV appears across R1 at room temperature, the voltage at the top of R1 is 60mV(1 + R2/R1). This voltage is proportional to absolute temperature (PTAT), and is referred to as V_{PTAT}.

The output voltage of the sensor element is 2 VBEs below V_{PTAT}. By proper scaling, the two VBEs can subtract 273°K from V_{PTAT} , thereby yielding an output that is proportional to °C. The graph illustrates how V_{PTAT}, $2xV_{BE}$, and V_{PTAT} $-2V_{BE}$ vary with temperature.

In contrast with thermistors, this technique produces a sensor that is inherently relatively linear. Internal trims and curvature correction for residual nonlinearities are normally included in this sort of circuit, but are not shown in this simplified schematic.

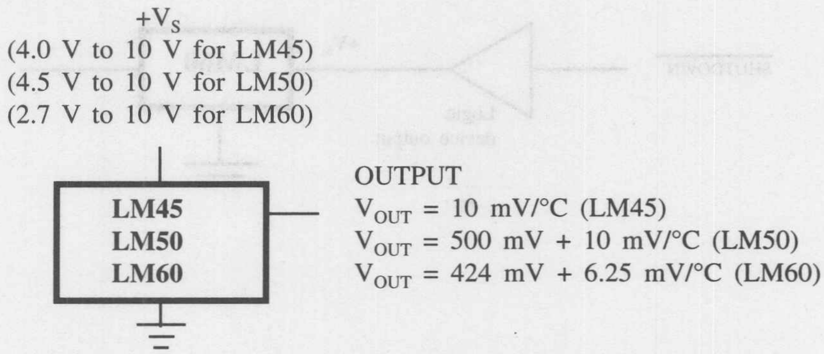
Semiconductor vs Thermistor



48

These curves compare the temperature-to-voltage transfer functions of silicon temperature sensors with that of an NTC thermistor. Thermistor nonlinearities can be corrected to some extent with lookup tables, but the inherent linearity of silicon sensors greatly simplifies system design.

LM45 / LM50 / LM60
Precision Centigrade Temperature Sensors in SOT-23 Package

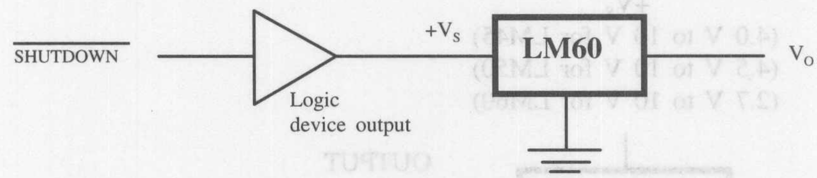


49

National Semiconductor offers several analog-output temperature sensors. The newest of these are the LM45, LM50, and LM60. A Selection Guide showing all of our temperature sensors can be found at the end of this section.

The LM45 is similar to the industry-standard LM35. They both have a convenient 10mV/°C output, with 250mV representing 25°C and 1.25V representing 125°C. The LM50 is similar to the LM45 with a 500mV positive offset. This will allow measurement of temperatures below 0°C (down to -40°C) without the need to connect the LM50 output load to a negative supply voltage. The LM60 has a lower output (6.25mV/°C) with a 424mV positive offset, and is designed for use in systems with power supply voltages as low as 2.7V. All three of these sensors are available in a 3-lead SOT-23 surface mount TinyPak package.

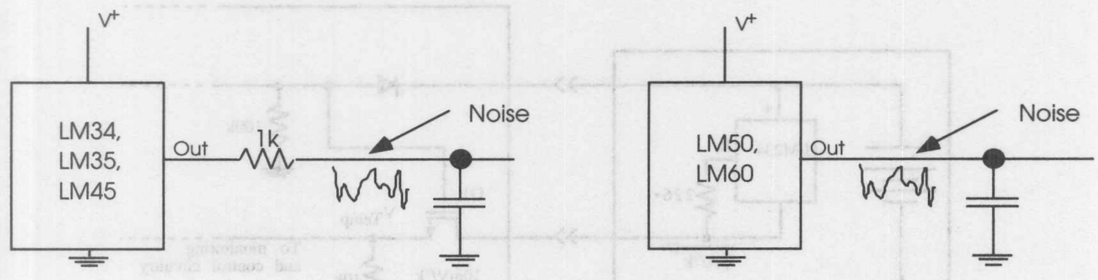
2.7V Temperature Sensor with "Shutdown"



50

A low-voltage, low-power sensor like the LM60 can be run directly from a logic gate, effectively imlementing a logic "shutdown" function for portable applications. The LM60 requires just 2.7V and 110µA to operate, so any logic gate that can provide that output voltage and current can power the sensor and shut it down when it is not needed.

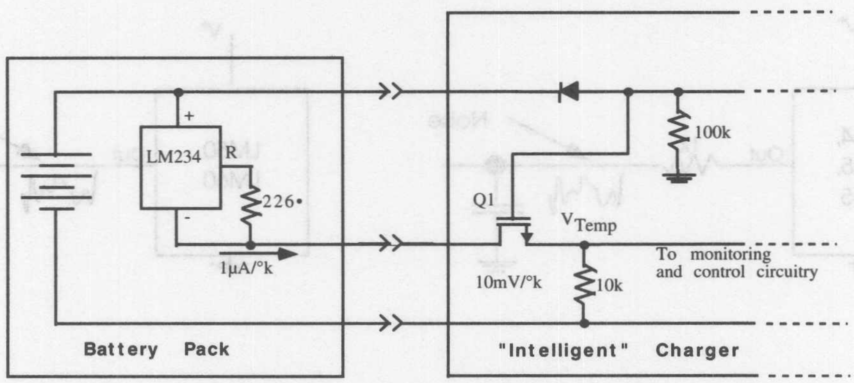
Noise Filtering



51

If your temperature sensor output must travel over long leads in a noisy environment, the analog signal may be corrupted by noise pickup. The conventional solution to this problem is to add a capacitor to the other end of the wire to low-pass filter the noise. The LM34, 35, and 45 don't like driving large capacitors directly, so it's a good idea to connect a 1k resistor to the sensor end of the cable to isolate the sensor's output stage from the capacitor. The LM50 and 60 have internal isolation resistors and don't need the external resistor.

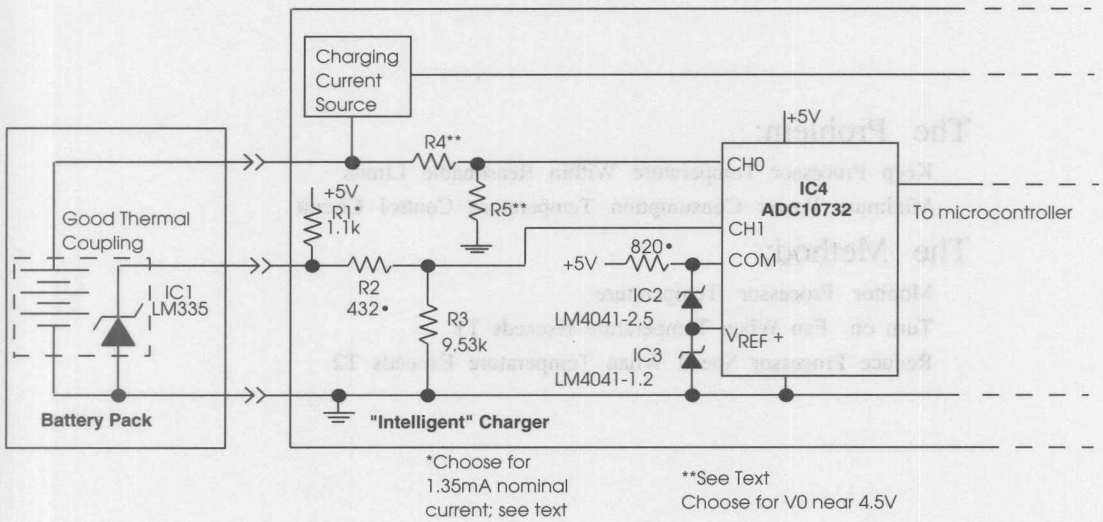
Battery Pack Temperature Sensor Can't Discharge Battery



52

This drawing shows a temperature sensor housed in a battery pack for charge control and safety enhancement. The LM234 produces an output current that is proportional to absolute temperature (1µA/K). This current can be converted to a voltage by connecting the LM234's output to an external resistor, which is located in the host system, or in the battery charger, as shown here. With a 10k• resistor, V_{TEMP} is 10mV/K. By using an external FET to break the current path, current drain by the sensor drops to zero when temperature is not being monitored. Sensor current drain also drops to zero when the battery is unplugged from the charger, or when it is plugged into a charger that has no ac power, thus preventing accidental battery discharge.

Voltage-Output Battery Pack Sensor



Here's another battery pack sensor, this time with a voltage output. More of the smart charger circuitry is shown in this schematic as well. The sensor is the LM335, which behaves like a two-terminal shunt voltage reference. It derives its power from the charger, and therefore cannot discharge the battery pack.

R1 provides about 1mA to the LM335, and an additional 350μA (at 350K) to divider R2 and R3. IC3, a 1.225V LM4041 reference, provides the reference voltage to the ADC10732 10-bit analog-to-digital converter. The attenuation provided by R2 and R3 reduce the LM335's 10mV/K scale to 9.57mV/K, which results in temperature-to-digital sensitivity of 8LSBs per K. The second voltage reference, IC2, is stacked on IC3 to provide a 3.725V (389K, or 116°C) reference point against which the sensor output can be differentially measured. Because the ADC10732 can measure negative input voltage differentials, this allows battery pack temperature to be measured from below 0°C up to the temperature limit of the sensor (100°C to 200°C, depending on device type, package, and duration of high-temperature operation).

The ADC also looks at battery pack terminal voltage to terminate charging. R4 and R5 should be chosen to attenuate the terminal voltage to the vicinity of 4.5V at the ADC's input when charging is to be terminated.

High-Speed Processor Temperature Control

The Problem:

Keep Processor Temperature Within Reasonable Limits

Minimum Power Consumption Temperature Control Circuit

The Method:

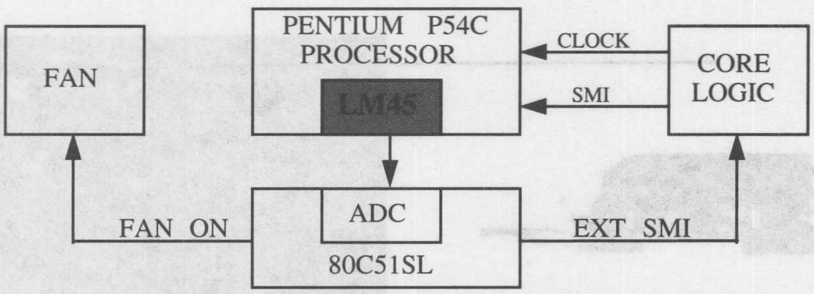
Monitor Processor Temperature

Turn on Fan When Temperature Exceeds T1

Reduce Processor Speed When Temperature Exceeds T2

Powerful, high speed processors, such as the Pentium®, PowerPC®, and Alpha®, add a lot of processing power to computers. They also consume a lot of power, and that can generate enough heat to impair reliability. To guard against this, manufacturers need to provide a program of thermal management. By monitoring the temperature of the processor, you can turn on a cooling fan when the temperature exceeds a predetermined danger level, then reduce the processor clock speed when a second, higher temperature is reached.

Pentium Active Thermal Feedback

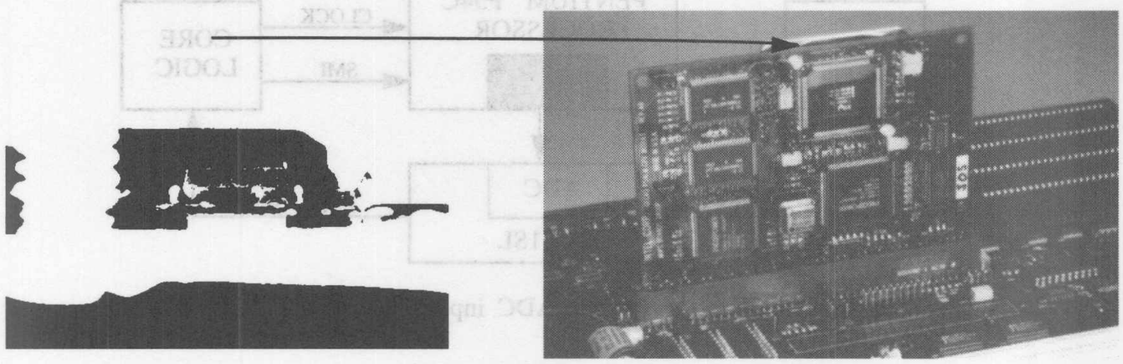


- LM45 Connected to the ADC input of the 80C51SL Microcontroller
- ADC Output Provides 2°C Resolution
- SMI (Interrupt) Generated Whenever Temperature Exceeds T_{HIGH} or Drops Below T_{LOW}

55

This block diagram shows how a temperature control system for a Pentium-based computer works. The LM45 temperature sensor is mounted in close proximity to the processor and its output voltage is monitored by the ADC in the on-board system housekeeping microcontroller. When the temperature exceeds the first temperature limit, the 80C51 turns on the fan to help carry heat away from the processor. When the processor temperature exceeds the next level, the 80C51 notifies the logic controlling the clock frequency to slow down the main processor clock and let the processor know that this has been done. The trigger temperatures will depend upon the thermodynamics of the system.

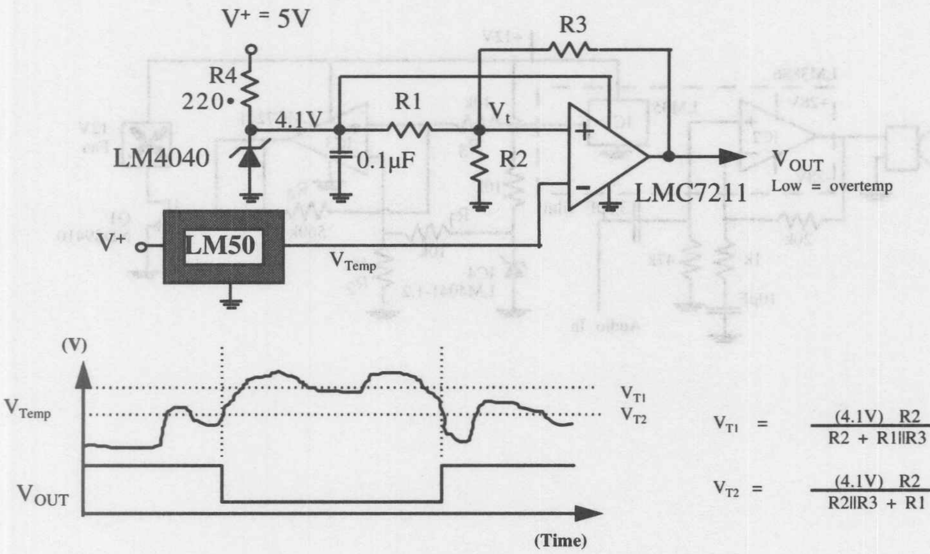
Thermal Monitoring



56

These photographs show the mounting of the LM45 in the Intel reference design. In this system, the processor is soldered to the daughterboard, and the heat sink is on the back side of the board. The LM45 is mounted between the heat sink and back side of the P54C printed circuit board.

Thermostat/Fan Controller



$$V_{T1} = \frac{(4.1V)\ R2}{R2 + R1 \| R3}$$

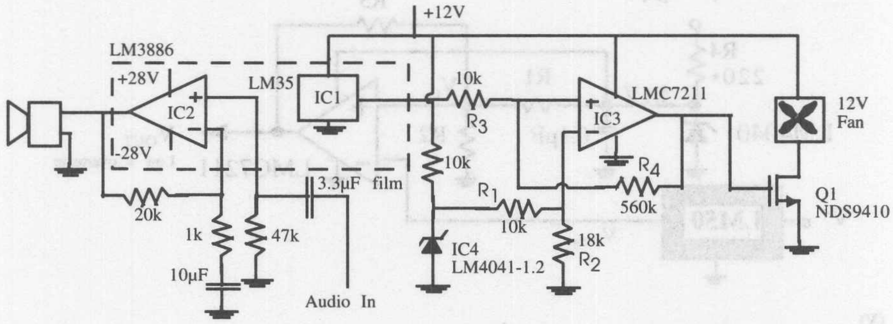
$$V_{T2} = \frac{(4.1V)\ R2}{R2 \| R3 + R1}$$

57

A simple fan controller like the one shown here can help to regulate temperature in systems with high-power processors. When small size is important, this circuit can be particularly useful, because it is built entirely from "tiny" components.

The LMC7211 comparator, available in a SOT23-5 package, is used to provide a thermostat output. The LM50 temperature sensor and LM4040 voltage reference are both in SOT-23 packages. R1, R2 and R3 set the hot (V_{T1}) and cold temperature (V_{T2}) trip points, as indicated by the formulas above. The LM4040 voltage reference is used for both a stable voltage source and as the supply voltage for the LMC7211, eliminating the need to route power to the amplifier. R4 is chosen so that there is no more than 15mA through the LM4040 when the LMC7211 output is high, and a minimum of 100µA through the LM4040 when the LMC7211 output is low.

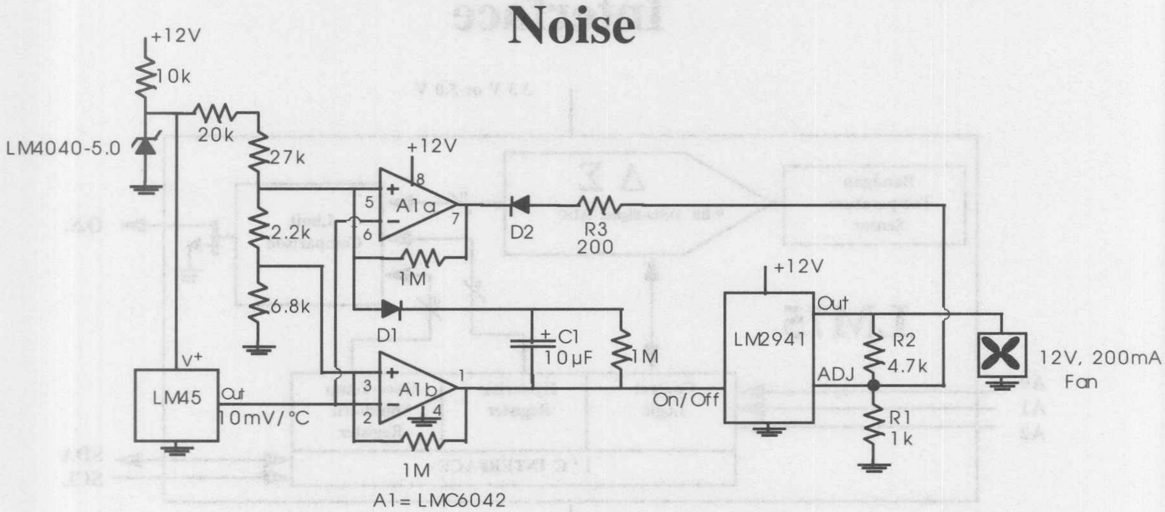
Audio Power Amp With Fan Controller



58

This circuit is a simple overtemperature detector for power devices. In this example, an audio power amplifier IC is bolted to a heat sink and an LM35 Celsius temperature sensor is glued to the heat sink near the power amplifier. To ensure that the sensing element is at the same temperature as the heat sink, the sensor's leads are also glued to the heat sink. The comparator's output goes low if the heat sink temperature rises above a threshold set by R1, R2, and the voltage reference. This fault detection output from the comparator now can be used to turn on a cooling fan. R3 and R4 provide hysteresis to prevent the fan from rapidly cycling on and off. The circuit as shown is designed to turn the fan on when heat sink temperature exceeds about 80°C, and to turn the fan off when the heat sink temperature falls below about 60°C.

2-Speed Fan Controller For Lower Noise



59

This fan controller helps solve the conflict between the cooling provided by a fan and the audible noise generated by the fan. The circuit keeps the cooling fan "off" when the system's temperature is sufficiently low, and operates the fan (at a moderately low speed) only if the system temperature becomes high enough to require forced air cooling. If the system heats up even more, the fan's speed increases to its maximum value.

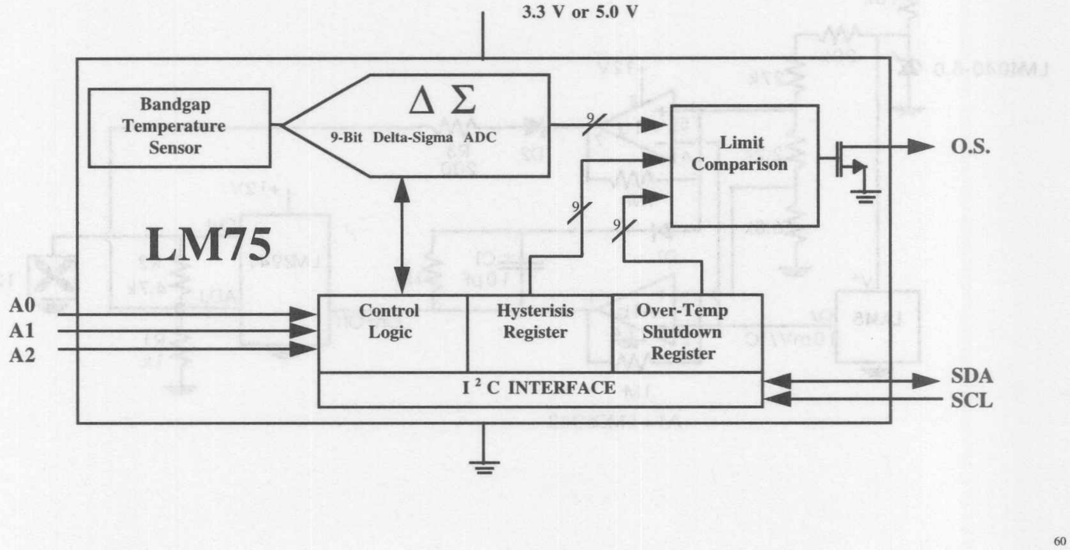
The LM45 temperature sensor's output voltage drives two amplifiers operating as comparators. The comparators' thresholds are set at 73°C (730mV) and 88°C (880mV) by the voltage reference and divider combination.

The LM2941 low-dropout voltage regulator supplies power to the fan. When the sensor is below 73°C, A1b's output holds the regulator's ON/OFF line high, which turns the regulator's output (and the fan) off. As temperature increases above 73°C, A1b's output goes low, turning the fan on. A1a's output will be high, which reverse-biases D2, so the regulator's output is

$$V_{OUT} = 1.275V\left(\frac{R1+R2}{R1}\right) \approx 7.3V$$

This will cause the fan to run at a relatively low speed. If the temperature continues to increase past 88°C, A1's output will go low, which places D2 and R3 in parallel with R1, causing the regulator output voltage to increase to nearly 12V. D1 and C1 provide a brief voltage boost when the fan first turns on at its lower speed. This results in a temporary 12V output from the regulator, which ensures that the fan starts up before its drive voltage drops to 7.3V. The 1M resistors between the amplifiers' outputs and non-inverting inputs provide approximately 9°C hysteresis to prevent the fan from cycling on and off too rapidly.

Sensor With Two-Wire Serial Interface



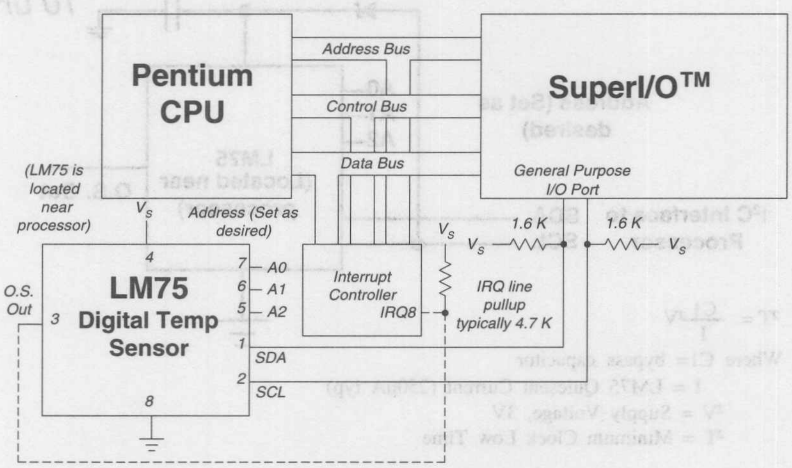
When a mostly-digital system (like a personal computer) needs a temperature monitoring function, the typical approach is to use an ADC to digitize the output voltage of an analog temperature sensor. By placing the ADC and the sensor on the same chip along with a two-wire serial interface, the size of the monitoring function can be reduced while improving functionality and accuracy.

The LM75 has a temperature sensor, a 9-bit (0.5LSB/°C) delta-sigma ADC, and an I² C® interface, which make it an ideal system-oriented solution for thermal watchdog applications. The host can program a temperature alarm threshold that triggers an output on the O.S. pin when the threshold is exceeded. A hysteresis temperature can also be programmed to provide a threshold at which the LM75 decides that the overtemperature condition is no longer present.

The delta-sigma ADC has low noise susceptibility, and a user-programmable "fault queue" gives further noise immunity by allowing the host to select the number of consecutive times the ADC must detect a threshold being exceeded before setting the O.S. pin.

Three address pins allow up to eight LM75s to share a single 2-wire bus.

Temperature Sensing on the PC Motherboard

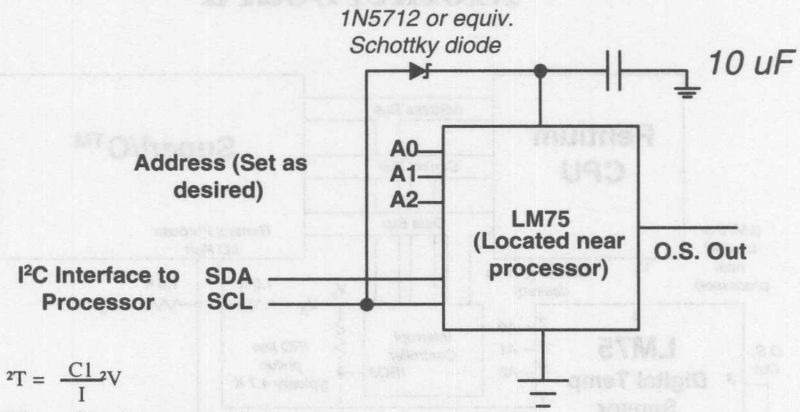


Here's an example of the LM75 in a PC thermal management application. The LM75 should be located near the CPU for good correlation between measured temperature and real CPU temperature. (See later in this section for more information on mounting).

The I^2 C bus interface in the PC is generated in a National SuperI/O. This bus can also be referred to as SMB, or System Management Bus. The SuperI/O provides several functions for computers including floppy disk and keyboard control, real-time clock, UARTs, infrared communications interface, several parallel ports and IDE interface. In this application one of the general-purpose parallel ports with open-collector outputs serves as the I^2 C I/O port. Up to 8 LM75s can reside on this bus, and each will be identified by an address set by connecting A0, A1, and A2 to ground or supply. The high bits of the I^2 C address are internally set to 1001.

The BIOS contains the software to implement the I^2 C communication which takes place between the CPU and the LM75 as well as any other I^2 C devices. The O.S. output of the LM75 makes it possible to detect overtemperature conditions without having to continuously query the LM75. The O.S. output is an open-drain output lending itself to connection to any open drain INT line such as the IRQ8 in a PC.

Clock-Powered LM75



$${}^{2}T = \frac{C1}{I}\,{}^{2}V$$

Where C1= bypass capacitor
I = LM75 Quiesent Current (250μA typ)
${}^{2}V$ = Supply Voltage, 3V
${}^{2}T$ = Minimum Clock Low Time

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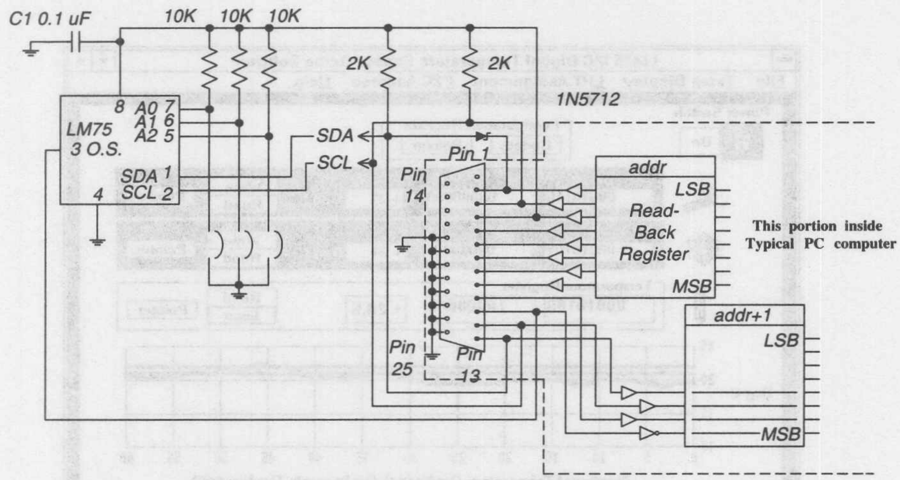
Because the I^2C interface rest state is high, the LM75 can derive power from the clock line. This eliminates either a power supply or wire. This sets a limit on the clock low time based on the value of the capacitor and current consumption of the LM75. The quiescent current of LM75 is 250 uA, although during communication it could reach 1 mA. Shutdown reduces typical quiescent current to 1 uA.

PC-Based Temperature Acquisition via Parallel Printer Port



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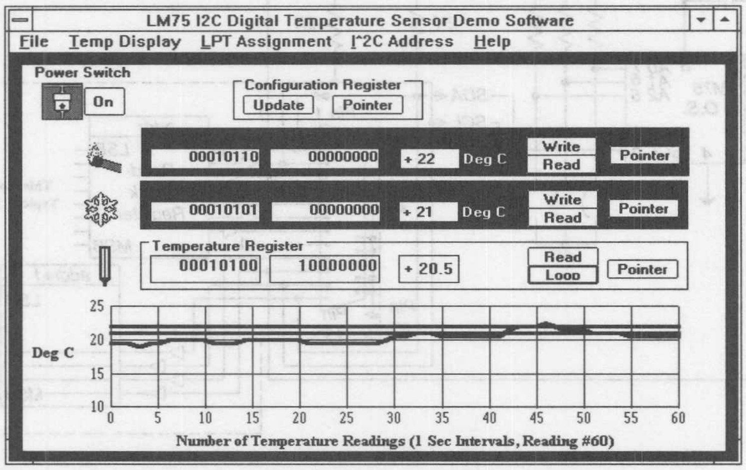
A demo board is available for the LM75 that allows PCs to acquire temperature data through the parallel printer port. The circuit is shown here. The LM75 gets its power from a line on the parallel printer port, although the bus-derived power of the previous slide could also be used.

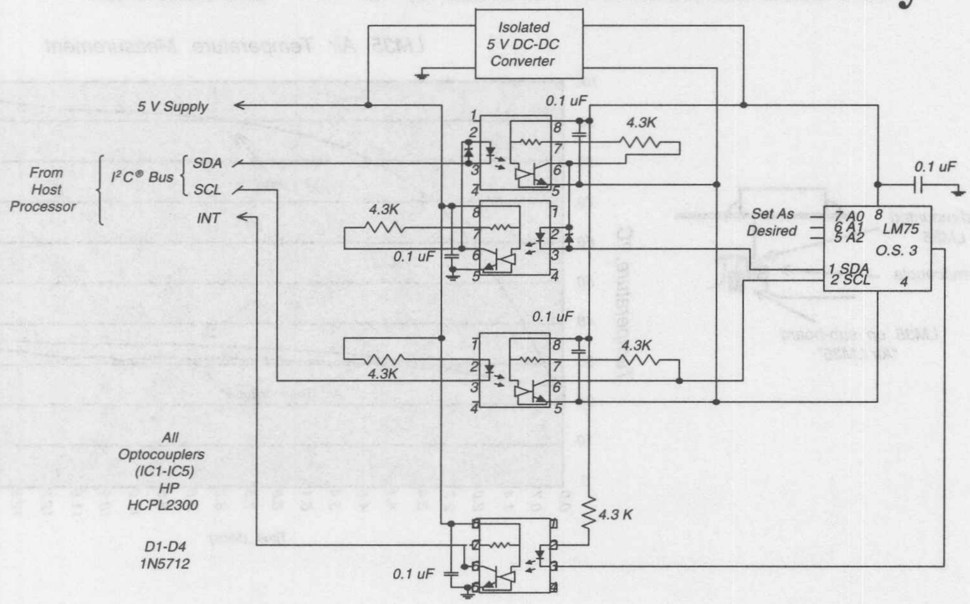
LM75 Digital Temperature Sensor Demo Software



64

This is the main "control panel" from the LM75 evaluation board software. The software allows you to read temperature, set thermostat thresholds, change internal configuration settings, and create temperature "strip charts".

Isolated PC-Based Thermometry



65

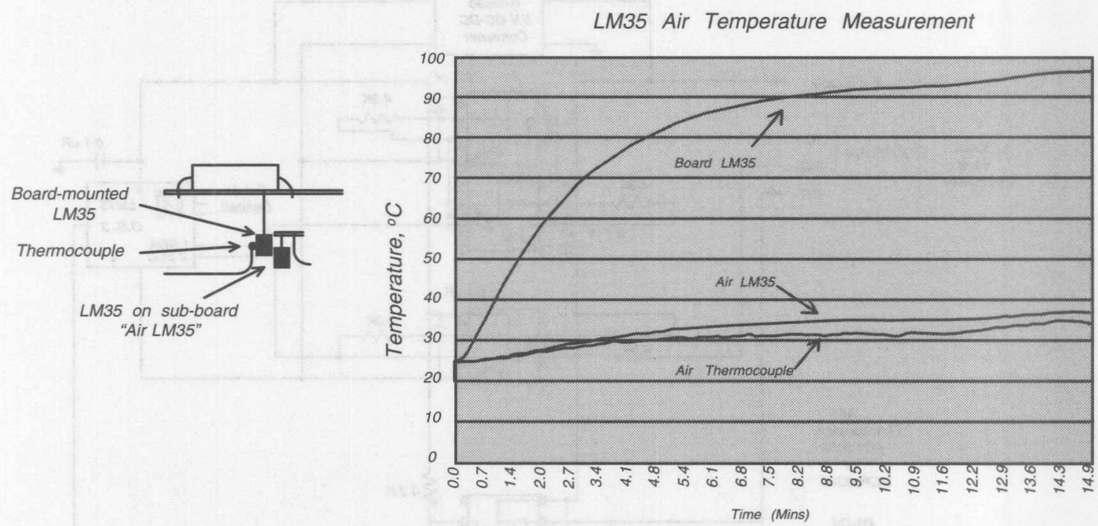
You can couple an LM75 digital output temperature sensor through the isolated I^2C interface shown earlier. Electrically isolating the sensor allows operation in situations exposed to high common-mode voltages; or could be useful in breaking ground loops.

Note that the SCL line is not bi-directional. The LM75 is a slave, and its SCL pin is an input only.

The O.S. optocoupler is optional and needed only if it is desired to monitor O.S.

Provide an isolated source of voltage, either by a DC-DC converter or a battery. The LM75 will operate from 3 V to 5 V, and typically requires 250 µA, while IC1 and IC3 require 7-10 mA each (the LEDs require about 700 µA, but only when active), for a total consumption of about 21 mA.

Where Does the Sensor Sense?



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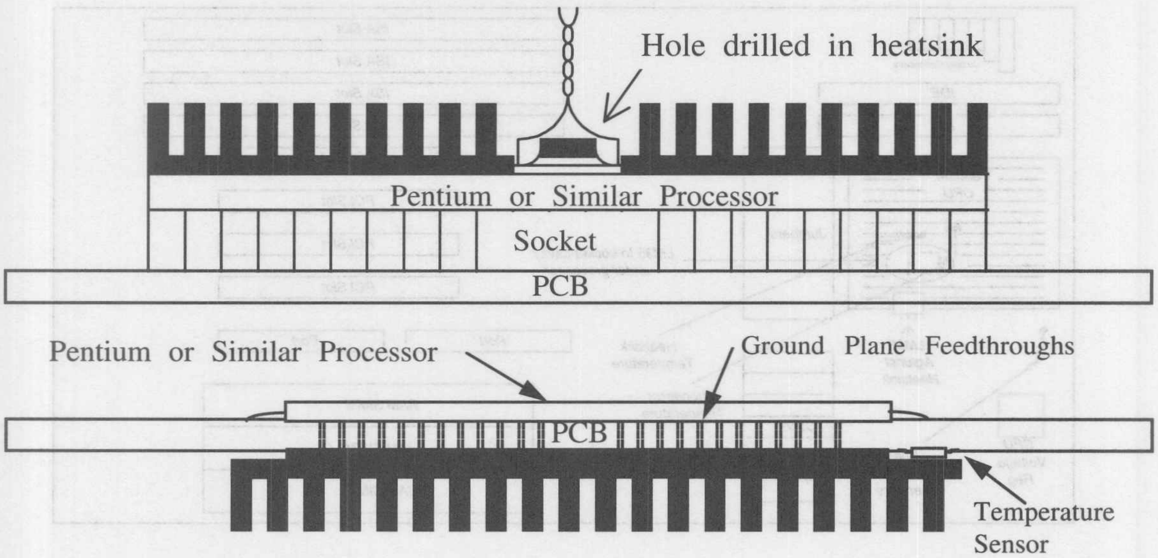
When you use a sensor to measure temperature, the largest source of error is rarely the sensor's inherent accuracy. Instead, the largest errors usually come from the sensor's temperature being different from that of the object being measured.

Most IC temperature sensors sense primarily through their leads because the leads have a much lower thermal resistance than the package. In fact, a common error is to use a sensor in a TO-92 package so that the package can protrude away from a board into an airstream. This does a poor job of measuring airstream temperature. Instead, it measures the temperature of the circuit board reasonably well, with a small error introduced due to the difference between board temperature and air temperature.

In this experiment, a circuit board was built with a heat generator (power resistors) on one side, and two LM35s in TO-92 packages on the other side. Power was applied to the resistors at t=0. One LM35 was mounted on the board, with its leads extended to place the package in the airstream. As you can see from the upper curve, its temperature was very nearly that of the board. The second LM35 was mounted on its own "sub-board"; and this entire arrangement is exposed to the aristream. The sub-board returns to the main board via small gauge wire. As the middle curve shows, this approach yielded a much lower temperature reading that was close to that of a thermocouple in the airstream.

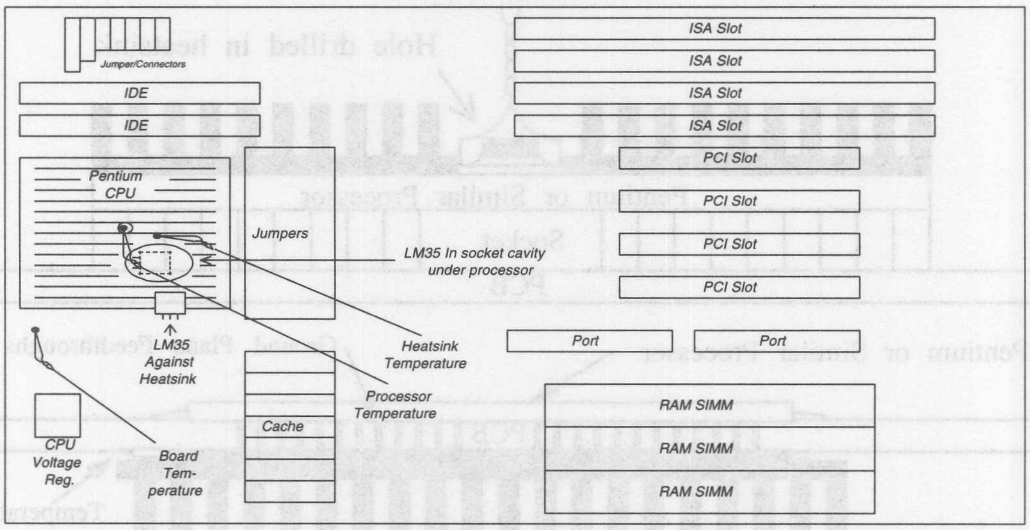
Pentium Temperature Measurement



The upper drawing shows an ideal way to measure the temperature of a Pentium or other high-performance processor. The processor chip is located just under the top center of the lid, so the sensor is mounted on a small sub-board in a cavity in the processor's heat sink. Because the sensor's leads are soldered to the sub-board, the sensor's temperature is very close to that of processor's case. Unfortunately, this installation is costly in mass production because of the sensor sub-board and the leads from the sub-board to the motherboard, which will require hand-assembly.

The lower drawing shows a more practical approach for systems that have the processor soldered to one side of the pcb and the heat sink on the other side. In this case, the sensor is just under the edge of the heat sink and has good thermal contact with the pcb, which is in good thermal contact with the processor, typically via the ground trace.

Practical Pentium Temperature Measurement



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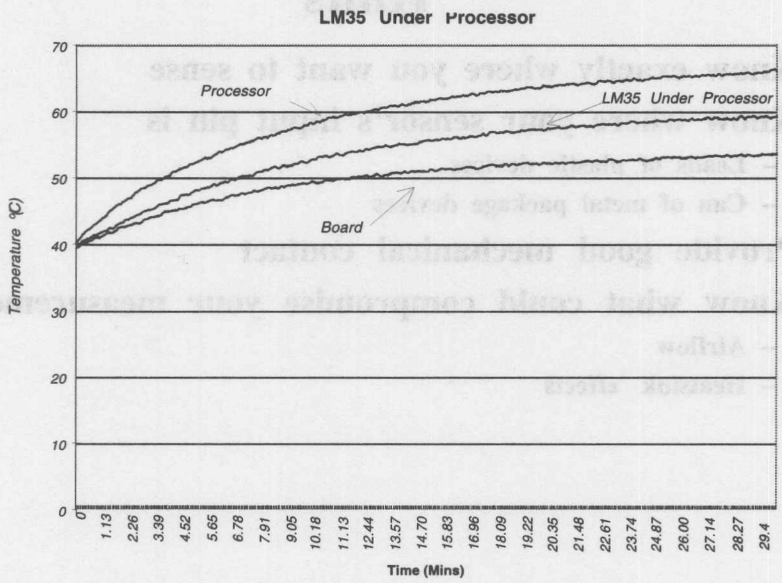
Testing has proven that one of the most practical places to install a processor sensor economically is in the socket cavity under the processor.(This assumes that the processor is in a socket as in the upper drawing on the previous page).

Experiments done with the typical board shown here indicate a good correlation between the under-socket readings and Pentium temperature. Mounting the sensor on the circuit board near the processor also correlates with processor temperature, but since a socketed processor is not in direct thermal contact with the circuit board, the sensor's temperature generally be several degrees lower than that of the processor. Some systems use multiple sensors located in various places on the board to monitor the temperature of other components that have potential for overheating.

Effect of Sensor Location



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These curves show how the sensor's location influences its temperature. The upper curve is the processor's case temperature, the middle curve is the temperature of a sensor below the processor in the socket cavity, and the bottom curve is the temperature of the circuit board near the processor. Note that the board temperature is several degrees below that of the processor, but it still provides some indication of the processor's temperature rise. When it is impossible or impractical to place the sensor under the processor, mounting it next to the processor can be a reasonable compromise if care is taken to correlate the sensor reading with the processor's actual temperature.

Temperature Sensor Installation Hints

- **Know exactly where you want to sense**
- **Know where your sensor's input pin is**
  - **Leads of plastic devices**
  - **Can of metal package devices**
- **Provide good mechanical contact**
- **Know what could compromise your measurement:**
  - **Airflow**
  - **Heatsink effects**

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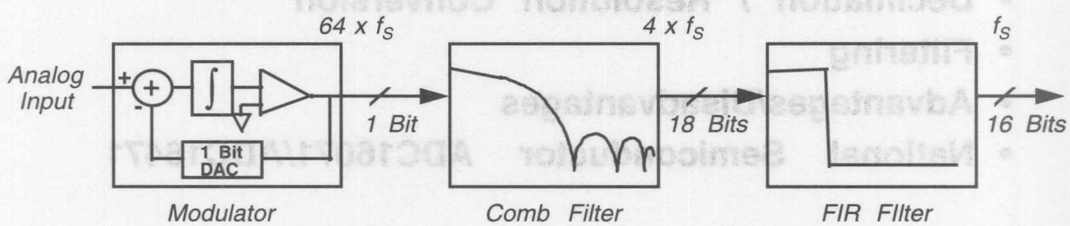
Appendix A: Oversampling Delta-Sigma Analog-to-Digital Converters

- Delta-Sigma Architecture Overview
- Noise Shaping
- Decimation / Resolution Conversion
- Filtering
- Advantages/Disadvantages
- National Semiconductor ADC16071/ADC16471

Oversampling Delta-Sigma ADC Block Diagram



f_S = Output Data Rate

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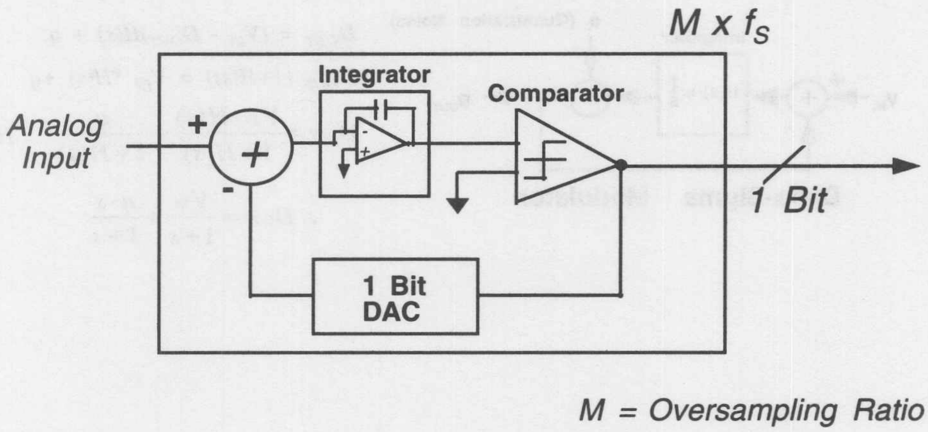
This is a block diagram of a simple oversampling delta-sigma (delta-sigma) ADC with a one bit modulator.

The Nyquist rate is the minimum rate, f_S, at which a quantizer must sample the input to prevent aliasing. The Nyquist rate is equal to twice the bandwidth of the analog input. When a quantizer samples at frequencies higher than the Nyquist rate, the digitized input can be faithfully converted back into a continuous analog signal. Delta-Sigma analog to digital converters sample the input signal at many times the Nyquist rate, thus the term "oversampling".

A delta-sigma converter consists of a delta-sigma modulator that is essentially a high speed, low resolution ADC , and a DSP block that trades time for resolution (ie: 64*f_s with 1 bit to f_s with 16 bits) and filters the output of the modulator. The DSP block typically consists of a comb filter, sometimes called a decimator, and an FIR filter that has a "brick wall" low pass characteristic.

Delta-Sigma Modulator



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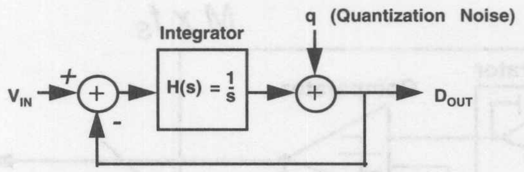
This is an example of a simple 1 bit modulator. The difference (delta) between the analog input and the comparator's previous output is integrated (sigma) in such a manner that the average of the digital output is equal to the analog input.

The ones and zeros of a modulator output represent the comparator's positive and negative full scale, respectively. For example, a modulator output of: 1,0,1,1,1,0,0,0,1,0 , represents an analog input half way between positive and negative full scale (5 out of a possible 10 ones).

Noise Shaping



Delta-Sigma Modulator

$$D_{OUT} = (V_{IN} - D_{OUT})H(s) + q$$

$$D_{OUT}\,(1+H(s)) = V_{IN} * H(s) + q$$

$$D_{OUT} = \frac{V_{IN} \cdot H(s)}{1+H(s)} + \frac{q}{1+H(s)} \qquad H(s) = \frac{1}{s}$$

$$\therefore D_{OUT} = \frac{V_{IN}}{1+s} + \frac{q \cdot s}{1+s}$$

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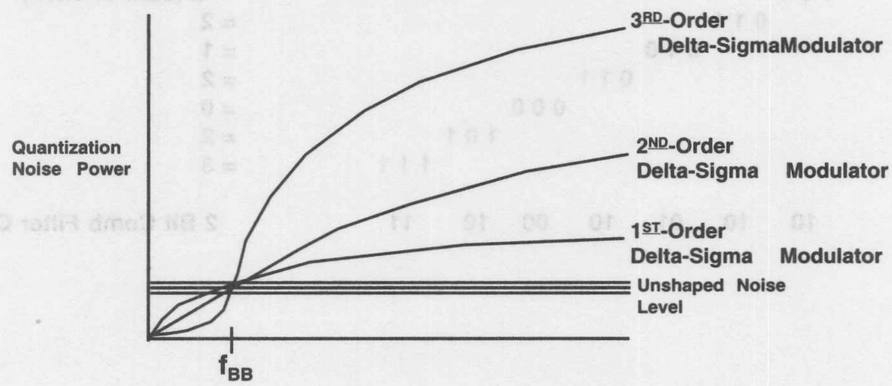
Because of the crude approximation made by the comparator of a delta-sigma modulator (only a 1 bit quantization), a large amount of quantization noise is introduced into the system.

If the comparator is treated as the addition of quantization noise to the output of the integrator, then the output of the modulator may be equated to the difference between the quantized output, D_{out}, and the analog input, V_{IN}, times the transfer function of the integrator, H(s), plus the quantization noise, q. From this transfer function, it can be shown that quantization is filtered through a high pass filter, while the input signal passes unattenuated at low frequencies ($f < f_{BB}$). This high-pass function "shifts" the quantization noise out of the baseband, f_{BB}.

Noise Shaping Characteristics



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The quantization noise added by the comparator is assumed to be white (spread out equally over the frequency spectrum). By increasing the order of a delta-sigma modulator (adding more integrators), the noise shaping effect is enhanced. These curves show how the flat quantization noise is "shaped" into 1st, 2nd, and 3rd order modulator characteristics.

Delta-sigma modulators further reduce the amount of quantization noise in the baseband by oversampling the input signal. The quantization noise is assumed to be spread out equally from DC up to the sample rate of the modulator. As the sampling rate is increased, so is the range over which the quantization noise is spread. The total noise does not decrease, but the density per frequency band does. With a first order modulator, the Signal to Noise Ratio can be shown to increase by 9dB with every octave increase the oversampling ratio.

By decreasing the amount of quantization noise in the input signal's baseband, via noise shaping and oversampling, delta-sigma modulators can be designed to achieve signal to noise ratios in excess of 100dB.

Decimating the Input Stream
(two bit output)

1 0 1 0 1 1 0 1 0 0 1 1 0 0 0 1 0 1 1 1 1 1	**1 Bit Modulator Output**
1 0 1	= 2 (sum of ones)
0 1 1	= 2
0 1 0	= 1
0 1 1	= 2
0 0 0	= 0
1 0 1	= 2
1 1 1	= 3
10 10 01 10 00 10 11	**2 Bit Comb Filter Output**

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One of the most difficult delta-sigma converter concepts to grasp is how a one bit modulator output is converted into an 8, 12 or 16 bit output word. To achieve resolutions of greater than one bit, the modulator output must be **decimated.** Through decimation, time may be traded for increased resolution. The decimation process illustrated here is similar to that of a simple comb filter. The modulator output is grouped into blocks of three one bit samples each. The ones in each block are summed, resulting in each block having a value of 0,1,2 or 3. Each sample of the comb filter output has twice the resolution and one third the data rate of the modulator output.

Decimating the Input Stream
(three bit output)

1 0 1 0 1 1 0 1 0 0 1 1 0 0 0 1 0 1 1 1 1 1	**1 Bit Modulator Output**
1 0 1 0 1 1 0	= 4 (sum of ones)
0 1 1 0 1 0 0	= 3
0 1 0 0 1 1 0	= 3
0 1 1 0 0 0 1	= 3
0 0 0 1 0 1 1	= 3
1 0 1 1 1 1 1	= 6
100 011 011 011 011 110	**3 Bit Comb Filter Output**

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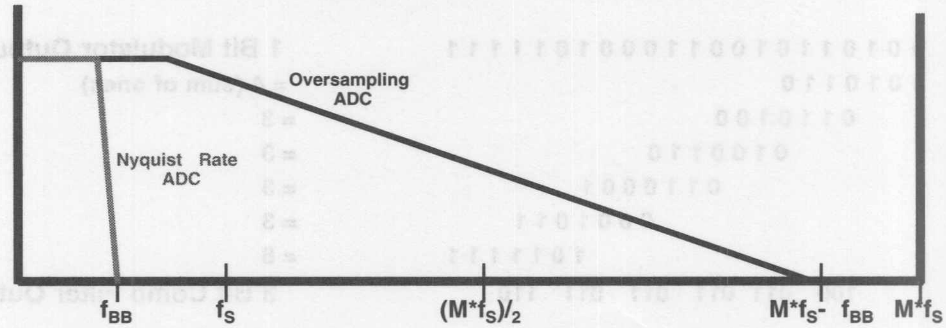
In this example, the modulator output is grouped into blocks of seven one-bit samples each. But this time, instead of shifting over the full width of one block, each grouping is only shifted over three bits. Summing the ones in each block results in a three bit value (0,1,2, ...7). This comb filter triples the resolution of the modulator output and decimates in time by a ratio of 3 to 1.

Note that the groupings containing 0110100, 01000110, 0110001, and 0001011 all result in the same sum: 3. This is evidence of the comb filter's low pass characteristic. Because of the comb filter's averaging, rapid changes in the modulator's output will be filtered.

Anti-Aliasing Filter Requirements
Oversampling ADCs vs. Nyquist Rate ADCs



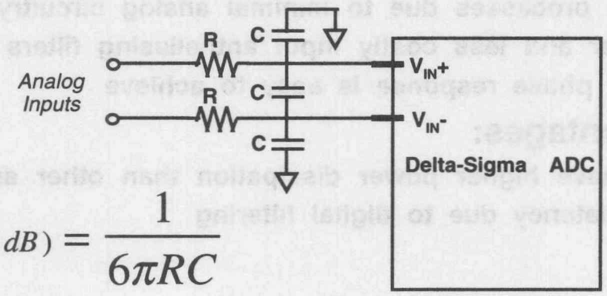
78

One of the biggest advantages of oversampling delta-sigma ADCs is their minimal requirement for anti-aliasing filters. To prevent aliasing with a Nyquist rate (non-oversampling) ADC, the analog anti-aliasing filter typically must have a flat response up to about .45 * fs and attenuate 60 dB or more by .5 * fs. Designing a filter with such a sharp cutoff can be difficult and expensive in both part count and board space. If phase distortion is a concern, the implementation can be even more difficult.

Aliasing will not occur provided no frequencies greater than one half the sampling rate are present at the input pins of an ADC. By sampling at many times the desired baseband, an oversampling ADC pushes out the point at which aliasing occurs. This enables the requirements for an anti-aliasing filter to be dramatically relaxed.

The critical point of attenuation for an oversampling ADC's anti-aliasing filter is typically pushed out even further because of on-chip digital filtering. The digital filters within an oversampling ADC typically cut off all frequencies above f_{BB}. Since aliased frequencies will be mirrored about half the sampling rate of the modulator, $(M^*f_s)/2$, any frequencies between $(M^*f_s)/2$ and M^*f_s will be aliased into the range between $(M^*f_s)/2$ and DC. Since all frequencies greater than the baseband (f_{BB}) will be filtered out by the on-chip digital filters, the only potentially damaging frequencies are those above M^*f_s - f_{BB} which will be aliased into the baseband. Thus the external anti-aliasing filter for an oversampling ADC need only cut off frequencies above M^*f_s - f_{BB}. With 64 times oversampling (M = 64), an anti-aliasing filter's critical point of attenuation is pushed out to 127 times (63.5^*f_s vs $.5^*f_s$) what it would need to be for a Nyquist rate converter!

Simple RC Anti-Aliasing Filter



$$f_{c(-3dB)} = \frac{1}{6\pi RC}$$

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This is an example of a simple RC low-pass anti-aliasing filter for the inputs to a delta-sigma ADC. If the filter's cutoff frequency, f_c, is set to around 4 times the desired baseband ($4*f_{BB}$) by choosing appropriate values for R and C, the filter should have very little effect on the signal in the baseband while attenuating all potentially damaging frequencies above $M*f_s - f_{BB}$.

Advantages/Disadvantages of Oversampling Delta-Sigma ADCs

- **Advantages:**
  - Easy and inexpensive to design ADCs with high accuracy on CMOS processes due to minimal analog circuitry.
  - Simpler and less costly input anti-aliasing filters
  - Linear phase response is easy to achieve
- **Disadvantages:**
  - May have higher power dissipation than other architectures
  - Long latency due to digital filtering

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Applications for Delta-Sigma ADCs

- **Good Applications for Delta-Sigma ADCs**
  - **Systems demanding low component count**
  - **Systems demanding linear phase response**
  - **Cost sensitive applications**
  - **High accuracy systems**
- **Difficult Applications for Delta-Sigma ADCs**
  - **Closed loop control/servo systems**
  - **Multiplexed Inputs**

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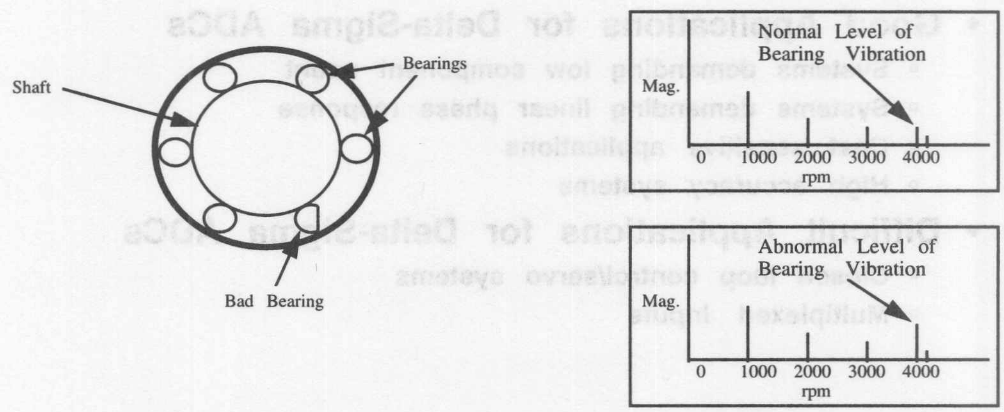
Because the digital processing within delta-sigma converters depends on many previous values, delta-sigma ADCs do not respond quickly to sudden changes or discontinuities in the input. Thus multiplexed inputs are not recommended for delta-sigma ADCs. This long latency time also prevents their use in all but the slowest closed-loop servo or control systems.

In addition to closed-loop frequency limitations, the delta-sigma ADC modulator can produce large amounts of dc offset. Some manufacturers compensate for this offset by using on-chip offset calibration. For delta-sigma ADCs without offset calibration, the dc offset can be compensated by a host system's processor. Therefore, applications demanding dc accuracy and high resolution can use non-calibrated delta-sigma ADCs with only a slight cost penalty of additional hardware and software.

National Semiconductor *Data Acquisition Applications*

Delta-Sigma ADCs In Action



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Vibration analysis is a good example of an application where National's delta-sigma ADCs could be used. A vibration analysis system monitors the frequency and magnitude of the vibrations coming from a mechanical system and compares them to a reference vibration analysis. The output of a sensor attached to the system is digitized and analyzed by a computer. If the "signature" of the noise is significantly different from the reference, this usually indicates that something is wrong with the system.

In the example given here, a rotating shaft (inside an electric motor, turbine generator, internal combustion engine, etc.) is supported by six round bearings. The circumference of the shaft is 3.78 times that of the bearings. This means that the bearings are rotating at 3.78 times the rate of the shaft. So, if the shaft is rotating at 1000 rpm, the bearings will be rotating at 3780 rpm.

If one of the bearings should wear out, as shown in the diagram, there will be an impulse generated every time the flat side of the bearing hits the shaft, causing an increase in the energy generated at the rotation rate of the bearing. The increased noise and/or vibration is picked up by a vibration analysis system, which recognizes that there is a problem with the bearings. The bearings can be replaced *before* catastrophic failure occurs.

Delta-sigma ADCs such as National's ADC16071 and ADC16471 are ideal for this sort of application because they don't require complex filters between the transducer and the ADC. They are also inherently very linear, providing excellent THD performance.